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PRELIMINARY STUDY OF MODULATION SYSTEMS FOR SATELLITE COMMUNICATION

Report No. 6R on
Contract No. NASw-495

S. Plotkin
June 1963

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HUGHES RESEARCH LAB~~ORATORIES~~
~~Malibu, California~~

a division of hughes aircraft company

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NASA CR-51993

Report [redacted]. 6R
Contract [redacted]. NASw-495

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SUMMARY

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A comparative study of power and bandwidth requirements for SSB, FM, and PCM modulation systems is given as related to multiple-access satellite communications. Composite power loading factors for multichannel speech are treated as a function of the number of channels. Actual system requirements for each of the three systems are then given with respect to an assumed reference application. Multiple-access considerations show that serious difficulties arise in both all-FM and all-PCM systems. Thus the concluding modulation system choice is SSB for the up link and wideband FM for the down link. It was further concluded that frequency division multiplex is preferable to time division multiplex. A theoretical comparison of the three modulation systems is derived which compares the peak transmitter power requirements as well as bandwidth. Results are then used to show the relative power requirements of equal bandwidth FM and PCM systems as a function of output S/N ratio. PCM is not beneficial unless very large channel S/N ratios are required, or a very small number of stations, and some serious practical problems are involved for large bit rates. Finally, even though the system bandwidth increases with the number of channels, the peak transmitter requirements for SSB and FM are nearly constant for 12 to 150 channels and are uncertain for less than 12 channels.

AUTHOR

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I. INTRODUCTION

This report covers the modulation system aspect of a broader study of multiple-access satellite communication systems. Its objective is to compare the probable suitability of the several attractive or widely discussed modulation methods for those satellite communication systems which permit simultaneous intercommunication between several (or perhaps many) earth stations sharing use of a common satellite repeater. Major attention will be devoted to the analysis and comparison of single sideband modulation (SSB), frequency modulation (FM), and pulse code modulation (PCM). Actually, certain of the modulation comparisons need not be restricted to use in multiple-access systems.

The controversy of active versus passive satellite repeaters has been practically resolved in favor of the active system as applied to world-wide nonmilitary communication systems. Rationale for this conclusion is based upon economics and recent technological advances such as Telstar which show the feasibility of reliable active repeaters. For the passive case, unless the area gain of the repeater equals the repeater gain plus antenna gains of the active system, an inordinate burden is placed on the earth station equipment. For reasons of this kind, this report envisages only active satellite repeaters.

One of the major purposes of this report is to bring out the constraints on the modulation system imposed by multiple-access operation. Another purpose is to present additional calculations regarding power and bandwidth requirements for various modulation schemes for a specific system. It is intended that the calculations can be easily modified to include various system changes as well as different system specifications from those assumed here.

Present designs of klystrons and traveling-wave tubes result in peak instantaneous power limitations 3 dB greater than the maximum average power rating of the tubes. Since transmitter costs are determined in large measure by the output tube, the total transmitter cost will be determined primarily by the instantaneous peak power. Thus this entire study is concerned with the instantaneous peak power requirements rather than the mean square or average power. It is noted that high power output tubes with instantaneous peak ratings considerably greater than 3 dB above their average values are being developed. If such tubes are successful they should reduce today's peak power cost penalty on the use of SSB.

A. Over-All Specifications

Perhaps the most important aspect of this report is the determination of essential features to be incorporated. Along this line, the following assumptions are given with a brief rationale for each.

1. A complete global system is envisaged which incorporates multiple access in the satellite repeater. It appears as though technological growth from a world-wide standpoint will be so rapid that some communication will be required from any one area of the earth to all other areas. For this purpose, multiple access would be at least the most straightforward solution as well as providing a backbone system for long term future demands.

2. Frequency sharing between satellite and ground microwave relay stations is mandatory. From an engineering judgment point of view, it appears as though communication satellites will have to operate in the already crowded 1 to 10 Gc region for a rather long period of time. It is pointed out that operation in the 15 to 20 Gc region was the subject of another report (see ref. 1). However, since initial systems, at least, will operate at lower frequencies, the most logical choice is to utilize the same or nearly the same bands as the ground microwave relays.

3. Bandwidth of the over-all system must be conserved, at least eventually. This point is rather obvious, but is spelled out to emphasize that both up-and-down links must be considered. Even if minimum bandwidth is not utilized in the initial systems, eventual reductions must be kept in mind to provide for increased traffic.

4. The final system must be within the financial capabilities of all countries. For a variety of political and economic reasons, it is unrealistic to conclude that the U.S. could or should finance the satellite communication systems for other countries. This is tantamount to saying that the system should be of minimum cost, but is actually a stronger statement. It implies that it is more important to plan a system which most nations can afford to use than to select a possibly less useful system primarily because it might become operational a year or so sooner. This comment relates to the modulation choice in that it must facilitate access by many stations, with an economic compromise between transmitter power, satellite life expectancy (i.e., simplicity), utilization of bandwidth, etc.

From the above, it should be clear that stationary satellites are advantageous. However, there are no conclusive technical reasons why nonstationary satellites in random or phased orbits could not be used to provide multiple-access operation. Their use merely introduces additional problems (such as Doppler, hand-over, etc.) and constraints (reduced one-hope coverage area, high fixed-cost of steerable antennas, etc.). Hence, this study of multiple-access modulation and multiplexing has not been restricted to any particular orbit system. As a consequence, it does not include modulation comparisons in relation to Doppler, hand-over synchronization, etc., although these could be significant considerations with nonstationary satellites.

B. Multiplex Methods

Before discussing the modulation system, the multiplexing must be considered. Essentially, there are three known ways of multiplexing a number of channels: FDM, frequency division multiplex; TDM, time division multiplex; and orthogonal or spread spectrum techniques.

Orthogonal or spread spectrum techniques make it difficult, but unnecessary, to distinguish between multiplexing and modulation. Signals are "coded" (modulated) redundantly, thus spreading their spectrum and imparting a noise-like character, but have properties of orthogonality which permit receivers to select desired signals from a heavy background of undesired (interfering) signals. Thus, many information channels can share use of the same broad RF bandwidth (i.e., be multiplexed) with their signals superimposed both in time and frequency. Demultiplexing is accomplished by the ability of receivers to recognize and to respond to the distinguishing characteristics (orthogonality) with which their desired signals were coded. Systems of this type may employ time-frequency address coding (see refs. 2 and 3), pseudo-random pulse coding (see ref. 4) or pseudo-random analog coding. All of these spread each narrowband information channel over the same wide transmission channel with noise-like characteristics. The proprietary and military interests in these techniques, and their early state of development, prevent giving a more detailed discussion and making an adequate evaluation of them.

From the multiple-access viewpoint, the most potentially attractive aspect of these techniques is that each receiving channel has a code to which it is responsive. Presumably (neglecting the possibility of simultaneous calls, synchronizing complications, etc.), a transmitter would have automatic access to any receiver merely by modulating with its code.

From the little that has been determined about such systems, they appear to have several serious limitations in relation to common-carrier type satellite communication. Foremost, perhaps, is the question as to whether their complexity can be reduced enough to become economically competitive with frequency division multiplexing. Second is the problem of relatively inefficient spectrum utilization because of a threshold effect. Suppose, for example, that the channel spectrum is spread by a factor of 1000 to a bandwidth which theoretically could carry 1000 such channels. With orthogonal modulation, not more than perhaps 100 channels could operate satisfactorily at the same time. Moreover, it is not clear how S/N is related to the number of channels in use, or whether such systems would degrade below the CCIR noise recommendations long before even these 100 channels were in use.

Jamming or interference protection is one of the primary advantages with orthogonal modulation, and is of prime importance in military systems but of only secondary consideration in commercial systems. In fact, as long as the output S/N ratio is satisfactory in the commercial system, it is undesirable to increase the bandwidth, transmitter power, or both, to obtain additional interference protection as would be the case in a military version. In view of this cursory analysis indicating that spread spectrum modulation will not at present provide any capability or advantages over the FDM or TDM systems in a commercial application, our attention will be confined to consideration of TDM and FDM only. Further consideration of orthogonal multiplexing as mentioned above may possibly be undertaken at a later date.

With regard to TDM, there are some formidable practical difficulties which have been recognized by the CCIR (see ref. 5). In fact, a direct quote seems applicable: "TDM has not, so far, been shown to be technically satisfactory for high-capacity radio-relay systems (300 telephone circuits or more)."

The problem, of course, results from the bits (pulses) becoming shorter as the number of channels is increased, thus complicating the switching and synchronizing problems. In a multiple-access satellite system these problems are further complicated by the unequal delays (distances) between the satellite and each of its earth stations whose pulses must arrive at the satellite within time-slots measured in nanoseconds. Theoretically, it is possible to correct for these delay differences leaving only nanosecond delay uncertainties to determine the time guard slots and then to reduce their number by transmission in pulse groups so that the cycle of transmission by all stations is slower. Though possible, such techniques introduce complications and thus increase costs. Also, such techniques have not yet been adequately reduced to practice. In short, it might be possible to achieve multiple access, with more than 300 channels, via time division multiplexing at the satellite of sequential transmissions from many earth stations, but this does not appear attractive nor competitive at the moment.

Another possibility would be to make combined use of TDM and FDM by using several 300-channel TDM groups on adjacent frequency sub-bands. In effect, the satellite would then have several repeaters, one for each TDM group, and this would introduce the multiple-access complications which were discussed in Sec. V of ref. 6.

It is noted that FDM is relatively straightforward, presents no serious problems for a multitude of channels, and is ideally suited for multiple access. Because TDM presents some serious problems for which solutions do not appear to be imminent, the conclusion is to consider FDM only. However, should TDM become feasible and advantageous in the future, it could be incorporated and might make PCM or delta modulation more attractive.

Having settled on the use of FDM, there remains the need to consider briefly how random access may best be achieved by many stations, large and small, in relation to the modulation method chosen.

The utility or worth of a random-access system increases with the number of its stations in that more population centers are accessible more directly. The number of stations in a one-hop system will depend heavily upon the economic feasibility of small stations, as was discussed in Sec. IX of ref. 6. Hence, the modulation system should be one which does not unduly penalize or limit the number of small stations and does not complicate the satellite excessively. Ideally, all earth transmissions should combine at the satellite into a composite FDM signal, which the satellite retransmits, perhaps with different modulation. Each station then receives and detects this composit of all channels but selects (filters out) only its own channels. Clearly, there are different difficulties in combining the transmissions at the satellite, depending on the modulation system chosen. Some of these difficulties, plus frequency-sharing aspects, are related to the possible presence of strong "carrier spikes," as will be discussed next.

Everyone is familiar with the carrier-beat interference conditions encountered with simple amplitude modulation (AM) as used in HF communication and for RF broadcasting. It is largely for this reason that AM need not be considered for satellite communication. FDM-FM has a carrier spike, the amplitude of which depends upon the channel loading and frequency deviation. With heavy loading and wide deviation, this spike may be negligible. With no channel loading, the entire transmitter power is radiated as a single-frequency carrier, unless carrier dispersal techniques are employed. Usual PCM transmission also involves a carrier frequency and both sidebands, but the carrier spike may be less sharp than with AM or FM. FDM-SSB is (or should be) carrier-free. Pilot tones may be transmitted to establish reference frequencies or for other purposes, but their amplitude can be relatively low and their frequency often can be chosen to avoid harmful beat-frequencies.

C. Systems to be Considered

Prior studies (refs. 7-10) have all considered one or more of only three modulation systems: SSB, FM or PCM. Power and bandwidth requirements for other types of modulation systems are well known so that a detailed comparison here is not deemed warranted. Additionally, multiple-access aspects of other systems are similar, in general, to those with either FM or PCM. One may add, however, that ΔM, delta modulation, is beginning to be recognized and might well be further evaluated in comparison with PCM. However, in this report only PCM is considered. The three modulation systems will be compared first by calculating power and bandwidth requirements for each case, followed by more general relative comparison of their bandwidths, peak powers, and information efficiencies. Figure 1 shows the general block diagram of the entire satellite system in which the master control satellite relaying (when necessary) and frequency synchronizing are not shown. Also, the encoding and decoding for PCM are contined in the modulator and demodulator respectively.

It is pointed out that PM and FM are so very similar that many papers do not distinguish between the two. In this report FM will be considered with emphasis and de-emphasis in order to equalize the (S/N) ratio in all channels. Thus, the higher frequency baseband channels will actually be transmitted PM while the lower ones will be FM.

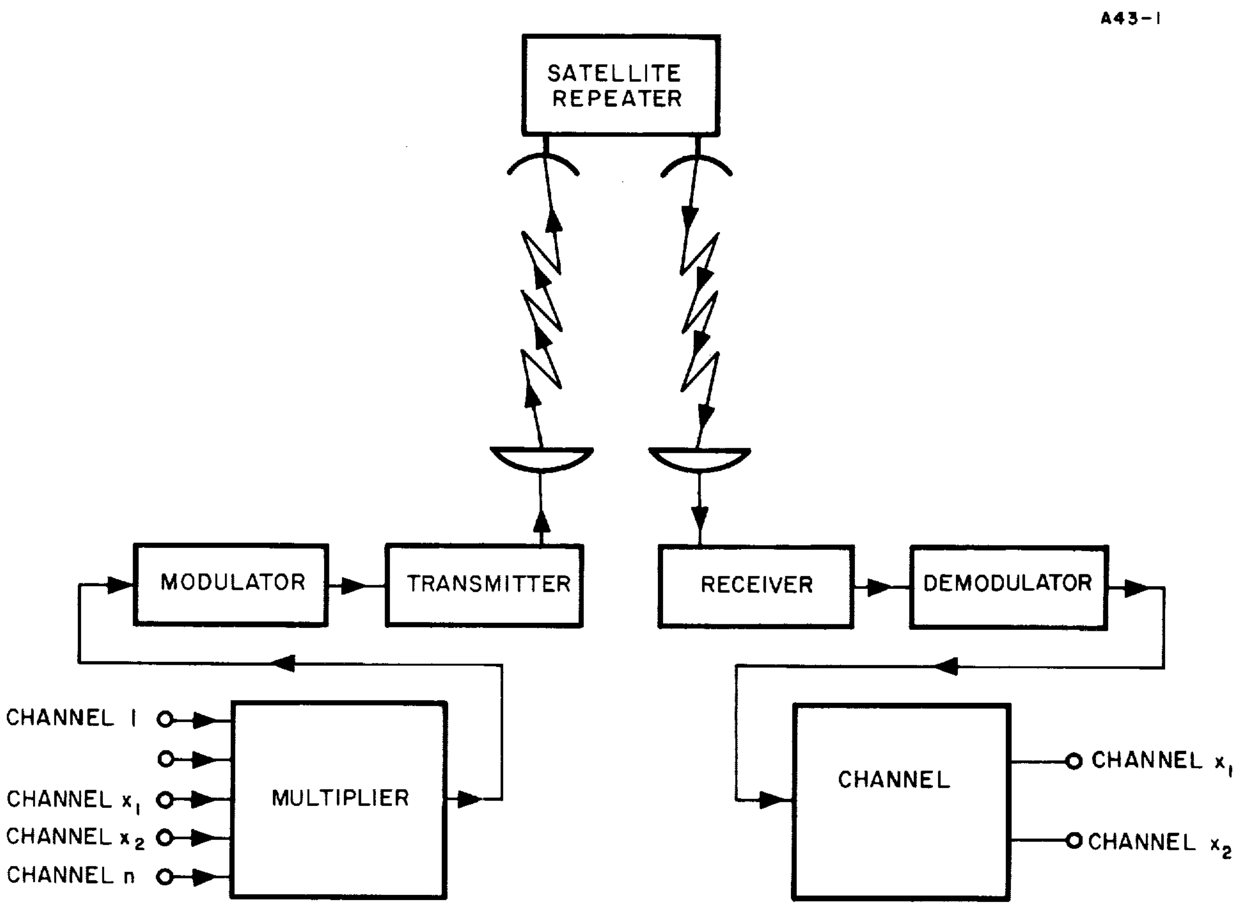

Fig. 1. Block diagram of a satellite communication system.

II. CHANNEL LOADING FACTORS

A clear understanding of the relation of average and peak signal power to the number of channels is of basic importance to much of the subsequent analyses and comparisons of modulation systems. The ratio of the "load" or total power in n channels to the reference "test tone" power in one channel is termed a loading factor. The standard channel test tone is an 800 cycle sinusoid having an average power of 1 mW at a point of zero relative level, such as at a toll test board. Elsewhere in a system the average power of this test tone may be attenuated or amplified but it still forms a useful reference for other powers such as those of signal or noise.

For example, if channel average noise power were 10^{-10} W (unweighted) at a point where the test tone had been attenuated 30 dB to 10^{-6} W, average, the test tone to noise ratio would be 10^{+4}, or 40 dB. After amplifying both back to zero relative level (1 mW test tone), the average noise power would still be 40 dB less than the milliwatt test tone, or at -40 dBm. Hence, the corresponding noise power anywhere in a system (assumed not to generate additional noise) can be specified as -40 dBm0, which means "minus 40 dB relative to a test tone whose average power at a point of zero relative level is 1 mW." One sees that test tone to noise ratios are more precise and useful than voice signal to noise ratios because even the average power of voice signals differ as much as 25 dB from one talker to another.

Holbrook and Dixon (see ref. 11) showed that the peak to average power ratio for n_a active voice channels* and for peaks exceeded not more than X percent of the time, decreased with increasing n_a or X. For high numbers of channels, these ratios approached those for random (gaussian) noise, about 10.2 dB for peaks exceeded 0.1% of the time. For one voice channel, the corresponding ratio is 18.4 dB. In an n channel system, studies have shown that each channel is "active" not more than 25% of the time during the busiest hours. The activity factor k is the maximum fraction of the n channels which probably would be active as much as 1% of the time during busy hours. Clearly $k \rightarrow 0.25$ as $n \rightarrow \infty$. However, a single channel would certainly be active more than 1% of a busy hour. Hence, k decreases from unity, for $n = 1$, toward 0.25 as $n \rightarrow \infty$. Figure 2 shows the Holbrook and Dixon peak to average power ratios in dB, replotted against the total number of channels, n, rather than against the number of active channels, $n_a = kn$. These ratios apply during the "most active" 1% of busiest hours. The curves have been fitted with the following empirical equations given by eq. (1):

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*Reference 11 uses n for the number of active channels and N for the total number of channels in the system. In this study n_a is the number of active channels, n is the total number of channels, and N is always a noise power.

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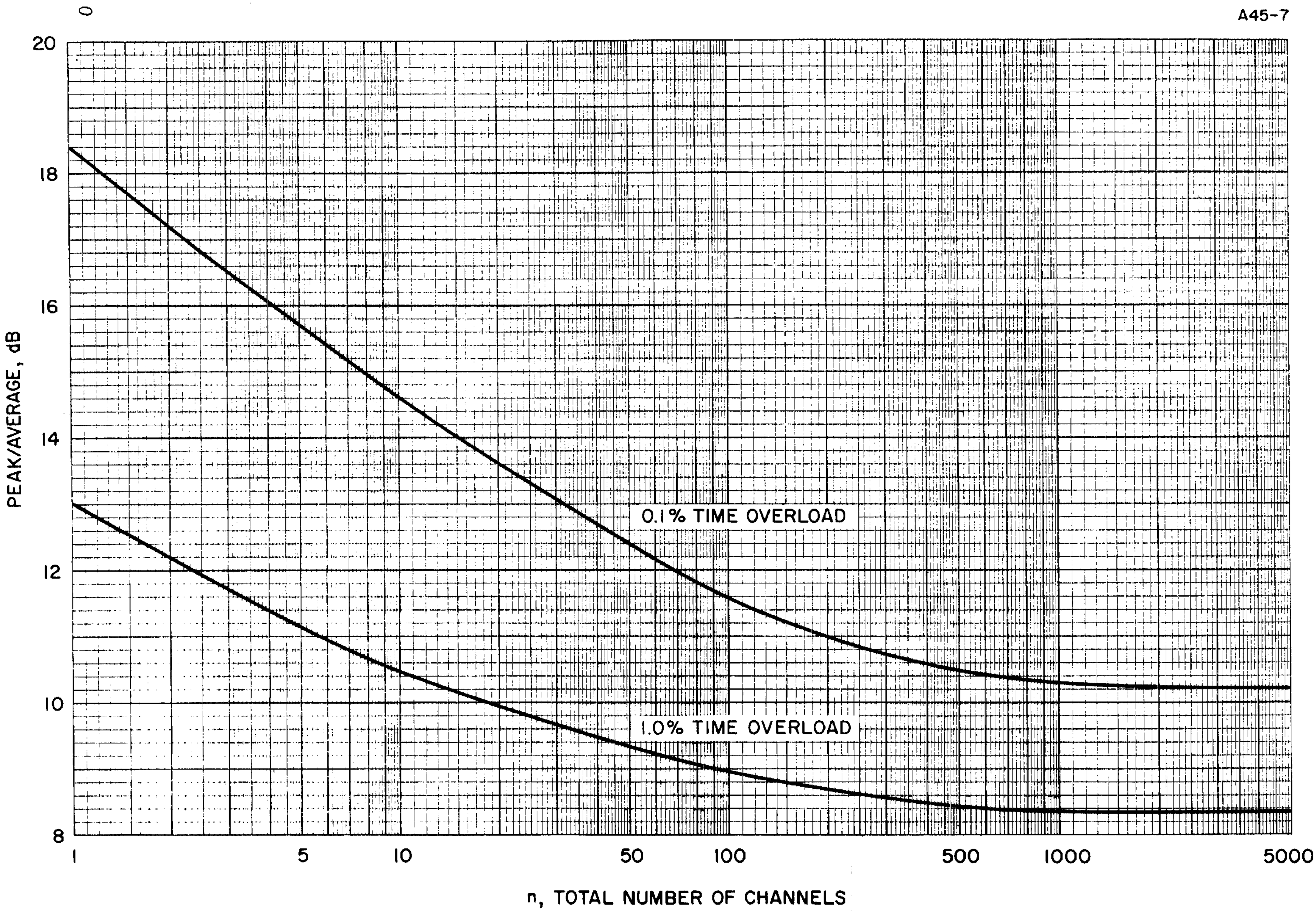

Fig. 2. Peak to average factor versus number of channels in the system.

$$p_f = 10.09 + 0.673\,(\log_{10} N - 3.55)^2;\ 10 \le N \le 10{,}000,\ \epsilon = 0.001$$
$$= 8.04 + 0.431\,(\log_{10} N - 3.39)^2;\ 10 \le N \le 10{,}000,\ \epsilon = 0.01 \qquad (1)$$

where

p_f = instantaneous peak to average power ratio factor, in dB

ϵ = fraction of time that p_f is exceeded.

To obtain the highest probable peak power for an n-channel system, one still needs to know the highest probable average powers to be increased by the above peak to average ratios. Even this presents problems because of the distribution of average powers among many talkers, because some channels may be used for tone (FDM) telegraphy at average powers higher than that of the average talker and because of the probable use of pilot tones.

The CCIR has adopted the following channel loading factors*, ℓ_f, for expressing n-channel average signal power during busy hours relative to 1 mW at a point of zero relative level as given by eq. (2):

$$\ell_f = -15 + 10 \log n \text{ dBm0, for } n > 240$$
$$= -1 + 4 \log n \text{ dBm0, for } 60 < n > 240, \qquad (2)$$
$$\text{and provisionally, for } 12 < n < 60\ .$$

These CCIR (average power) channel loading factors are shown by the two straight lines in Fig. 3. The lower line has been dashed for range $12 < N < 60$ to show that the loading factor is only a provisional one within this range. Additionally, this line has been extrapolated to -1 dBm0 at n = 1, lightly dashed to show the questionable nature of such an extrapolation. The behavior of fewer than 12 channels is statistically uncertain. In the extreme, a single channel could be used by a weak talker, averaging -25 dBm0, by a near-average talker at about - 10 dBm0, by a loud talker at 0 dBm0 or by multiplex telegraphy at an average power

*Contained as notes to CCIR recommendations relative to noise in FDM hypothetical reference circuits. For example, see Note 8 to Recommendation 287, (Los Angeles, 1959) applicable to most microwave relay systems, or Note 4 to the new recommendation applicable to satellite communication systems, adopted as Document 2173 in Geneva, 1963.

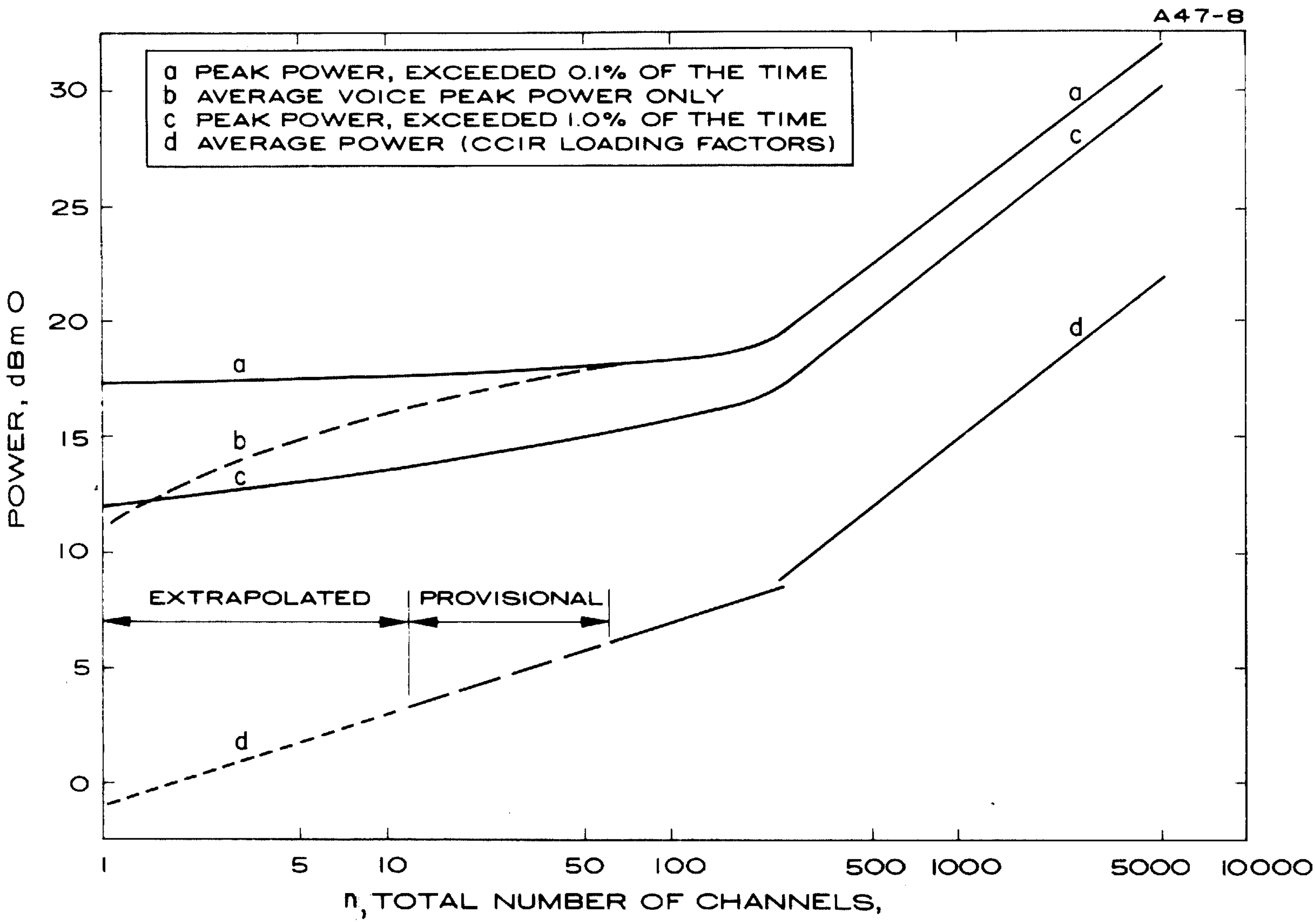

Fig. 3. Power loading as a function of the number of channels.

in excess of 0 dBm0. Altogether, these CCIR channel loading factors are empirical relations, simple, convenient and universally used, but possibly tending to be more conservative than accurate.

The upper solid curves in Fig. 3 were obtained by adding the peak to average power ratios of Fig. 2 to the CCIR channel loading factor average power curve. Thus, subject to reservations about the legitimacy of extrapolating to average powers for $n < 12$, these upper solid curves show design values of peak power in dBm0 versus n, for peaks exceeded 0.1% or 1.0% of the time. Values from these curves will be referred to as peak loading factors, $(p_f + \ell_f)$.

A significant and often unfortunate aspect of these peak loading factor curves is that they indicate nearly constant peak power for small numbers of channels. One would expect that the total peak power from voice channels would not increase with the number of channels as rapidly as their total average power for small numbers of channels. Until there were many channels, it would be relatively seldom that peaks from different channels would coincide. Nonetheless, there are reasons to believe that, for small numbers of channels, these peak loading factor curves are too pessimistic. For example, a more optimistic peak loading factor for small numbers of channels is shown by the dashed curve which diverges from the 0.1% time-overload curve. This curve was obtained by using the shape of Holbrook and Dixon's peak power curve, B in their Fig. 8 of ref. 11 and matching it to the 0.1% time-overload curve at $n = 60$ channels. The $n = 1$ peak power is seen to be some 6 dB lower. It should be recognized, however, that this dashed curve assumes voice channels only, with no pilot tones.

Thus it has been shown in Fig. 3 that by combining the Holbrook and Dixon peak-to-average power ratios with the CCIR average power loading factors, the relative peak power (dBm0) of frequency division multiplexed voice channels is relatively constant for 12 to 120 channels, increasing ~ 1.5 dB for $\epsilon = 0.01$ and only ~ 0.9 dB for $\epsilon = 0.001$. For less than 12 channels the relative peak power is uncertain, partly because the CCIR loading factor applies only for 12 or more channels but basically because it can be influenced so much by one or more loud talkers or by the use of tone telegraph. For more than 240 channels, both peak and average power increase at 3 dB per octave, and in the same ratio as for random noise.

These results probably are conservative, even pessimistic, for several reasons. Basically, they are a consequence of the peaky nature of speech and of the wide distribution of talker volumes, in addition to the empirical nature of the CCIR loading factor, the possible use of tone telegraphy, etc. The seriousness of these results, in relation to the small stations of a multiple-access system using SSB, lies in the fact that present microwave transmitters operate at constant power input, with peak power capability. The economics would improve if these transmitters could operate more nearly at the average power and draw more power for the peaks, more like a conventional class B or C amplifier. Such operation seems promising with new traveling wave tubes, having depressed collectors. Beyond this, major improvement of the signals should be possible. At the very least, automatic or manual gain control could reduce the spread of average powers between weak and load talkers. Ideally, compandors should reduce the peak-to-average power ratios of individual voice channels, as well as equalizing their average powers, so that the statistics of random noise would be applicable with fewer channels.

III. REFERENCE SYSTEM PARAMETERS

The CCIR has recommended* that the psophometrically weighted noise, averaged over any hour, in any channel of a one-hop (hypothetical reference) satellite communication circuit not exceed 10, 000 pW, measured at a point of zero relative level. This allowance is total for both intermodulation and thermal noise in both up and down paths. Thus, the thermal noise allowance for either path may be somewhere between 1, 000 and 5, 000 pW during the worst (i. e., "any") hour, or less during an average hour. Illustratively, we will allow 1, 000 pW, or -60 dBm0, psophometrically weighted for each path, which is equivalent to -57. 5 dBm0 flat weighted.

It will be seen that a straightforward SSB design for an earth to satellite link would require a transmitter of high peak power capability to carry even a few voice channels. Hence, it is assumed that compandors will be used to provide an effective noise reduction (S/N improvement) of 13.5 dB. This value seems reasonable, since considerably higher compandor improvements have been claimed (ref. 12). It is recognized however, that compandor noise improvement is somewhat subjective and is associated with its "hush-hush effect" of suppressing noise during quiet periods. Also, it has been the practice of common-carriers not to use compandors extensively. However, it will be seen that the economic importance of compandors in satellite communication justifies their use and justifies effort toward their improvement.

It should be noted that compandors would be used on each voice channel at earth stations, not in the satellite. The compandor advantage applies to both paths, to and from the satellite. Ideally, its effect is to suppress the total noise, hence the additive noise of both paths, by the postulated amount. Instead of requiring -57. 5 dBm0 noise for each path, we let it be -44 dBm0 by assuming a 13. 5 dB compandor improvement.

Present syllabic compandors require about 50 msec time constants in order to act upon syllables rather than instantaneous amplitudes. The question at this point is what, if any, effect this has upon the peak loading factor. Stated another way, do the peak amplitudes in speech have rise times faster than the compandor attack times? The general conclusion assumed here, and certainly a conservative conclusion, is that the peak loading factor is essentially unaffected by compandor action. Whether or not this conclusion is valid requires extensive tests beyond the scope of this report. Therefore, compandor improvement of 13. 5 dB is assumed and uncompanded peak loading factors will be used.

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*Document 2273 Xth Plenary Session, CCIR, Geneva, 1963.

Subsequent comparison calculations, for SSB, FM and PCM, will be made on a basis of the following system parameters which were chosen for illustrative purposes and partly as a compromise between typical "up" and "down" parameters. For example, 300°K has been used as the receiving system temperature both in the satellite and at earth stations. 60°K might be more typical for the latter, or 3000°K for the former. However, results based on 300°K can be scaled easily to these or other temperatures. Similarly the results can be scaled for different compandor improvements or noise allowances in any final system design.

TABLE I

ASSUMED SYSTEM PARAMETERS

Parameter	Symbol	Up	Down
Frequency, Gc	f	6.0	4.0
Max. path length (5° elevation) satute miles,	D	25,600	25,600
Receiving system noise temperature, degrees Kelvin	T	300	300
Satellite antenna gain (earth-subtending), dB,	Gs	19.5	19.5
Earth antenna gain (40 ft parabola), dB,	Gr	55.0	51.5
Miscellaneous losses and degradation margin, dB	L	10.0	10.0
Channel bandwidth in cps,	B_{ch},	4,000	4,000

As additional comments, a 60-ft parabola might have been a more typical choice for the earth antenna, increasing the above gains by 3.5 dB. In part, the 40-ft operation was chosen at a time when less expensive 30-ft antennas were being advocated and when full gain at 6 Gc from a 60-ft antenna seemed questionable. The 10 dB miscellaneous loss allowance includes beam-edge gain reduction, losses in diplexers, transmission lines, etc., possible polarization loss, miscellaneous increases in receiver noise temperature, etc. This 10 dB is believed to be reasonably conservative, considering that some allowance for rain or other atmospheric absorption has already been made by allocating only 1000 pW of psophometrically weighted noise to each link. The 4 kc channel width also was chosen for convenience, the CCIR channel width being only 3.1 kc. 4 kc is closer to the channel spacing, so that with n channels 4n kc can be taken as approximating the baseband. In effect, assuming 4 kc channels neglects the guard bands between channels; there would be 1.1 dB less noise in a 3.1 kc channel. However, CCIR basebands are somewhat greater than 4n kc as shown in Table II, which is taken from ref. 5 (See Tables 2 and 4).

TABLE II

BASEBAND CHARACTERISTICS

Telephone Channels, n	CCIR-CCITT Baseboard limits kc/s	f max for pre- and de-emphasis kc/s	4 n, kc/s
12	12-108	108	48
24	—	—	96
60	60-300	300	180
120	60-552	552	480
300	60-1364	1300	1200
600	60-2792	2660	2400
960	60-4287	4188	3840
1200	—	5686	4800
1800	60-8248	8204	7200

IV. REFERENCE SYSTEM COMPARISONS

A. SSB Calculations

The multiple-access aspects of FDM-SSB are particularly straight forward. Transmissions from each ground station are simply added together at the satellite receiver input to form a composite baseband of n channels, each occupying a separate 4 kc frequency slot. The only stringent requirement is that the frequency translation from baseband to the proper 4 kc slot at 6 Gc must be precise in order to avoid overlapping of channels. At the satellite the choice prevails of simply translating the composite input at 6 Gc to 4 Gc using SSB down, or converting to another modulation system at the satellite. However, only the variation in power with the number of channels is required for the fundamental considerations of this study.

With SSB detection, the RF bandwidth is the baseband, $n\ B_{ch} = 4.0\ n$ kc, for n channels. The noise in this bandwidth, at 300°K, is

$$P_n = 10 \text{ Log } (KTNB_{ch}) = 10 \log (1.38 \times 10^{-23} \times 300 \times 4000 \times n)$$

$$= -167.8 + 10 \log n \text{ dBW} \tag{3}$$

and the per channel noise is -167.8 dBW. This corresponds to -44 dBm0, the value previously calculated from 1000 pW (weighted) and the 13.5 dB compandor improvement. Hence, the 0dBm0 level at the receiver input per channel is

$$44 \text{ dB} - 167.8 \text{ dBW} = -123.8 \text{ dBW} .$$

Next, the path and transmission losses will be calculated, using the 6 Gc "up" frequency

$$A = 37 + 20 \log f_{(mc)} + 20 \log D_{(miles)} = \text{path loss}$$

$$= 37 + 75.5 + 88.2 = 200.7 \text{ dB} \tag{4}$$

$$L_t = A + L - G_R - G_S = \text{total transmission loss}$$

$$= 200.7 + 10 - 55 - 19.5 = 136.2 \text{ dB} \tag{5}$$

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Note that this transmission loss is independent of frequency because G_r and A increase equally with frequency. The transmitter's reference level, P_{TO} (for 0dBm0) must be greater than that at the receiver by the amount of this transmission loss, or P_{TO} = 136.2 dB - 123.8 dBW = 12.4 dBW, or 17.4 W. A channel test tone transmitted at this average power would reach a point of zero relative level with an average power of 1 mW, 0dBm0. Thus, to determine the instantaneous peak transmitter power which would not be exceeded more than 1%, or 0.1% of the time during busy hours, it is only necessary to add 12.4 dBW to the appropriate peak loading factor curve (taken in dB) of Fig. 3. This has been done for the SSB curves of Figs. 4 and 5, for comparison with FM and and PCM systems.

Of course, the same results can be obtained on a multichannel basis in completely algebraic form by making use of eqs. (1) (2), (3), and (5).

$$P_{T_{SSB}} = \text{peak instantaneous transmitter power for SSB}$$

$$= L_t + \left(\frac{S}{N}\right)_c + P_n \qquad (6)$$

where

$$\left(\frac{S}{N}\right)_c = \frac{\text{peak instantaneous transmitted power of the composite signal}}{\text{average noise power in the base bandwidth}}$$

$$= \left(\frac{S_{av}}{N}\right)_{ch} + l_f + p_f - 10 \log_{10} n \qquad (7)$$

$$\left(\frac{S_{av}}{N}\right)_{ch} = \frac{\text{average signal power of one channel}}{\text{average noise power in one channel baseband}}$$

For SSB the base bandwidth is equal to the base bandwidth which is different from the FM and PCM cases. It is noted that Pn contains a + 10 log n term which cancels with the -10 log n in eq. (7). Therefore, had CCIR bandwidths from Table II been used rather than nB_{ch}, they would have cancelled out leaving eq. (6) as a function of n only.

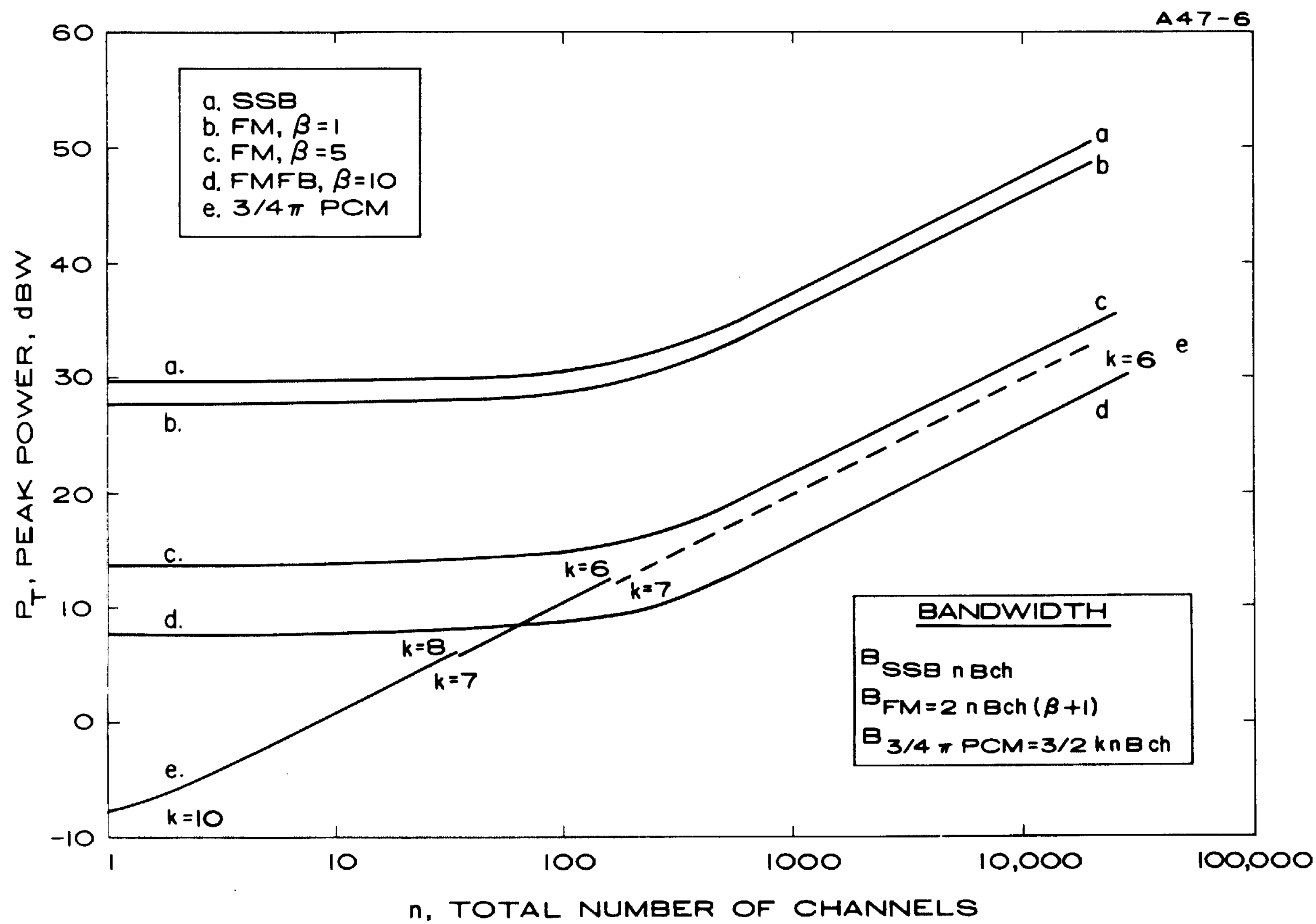

Fig. 4. Peak transmitter power, 0.1% time overload.

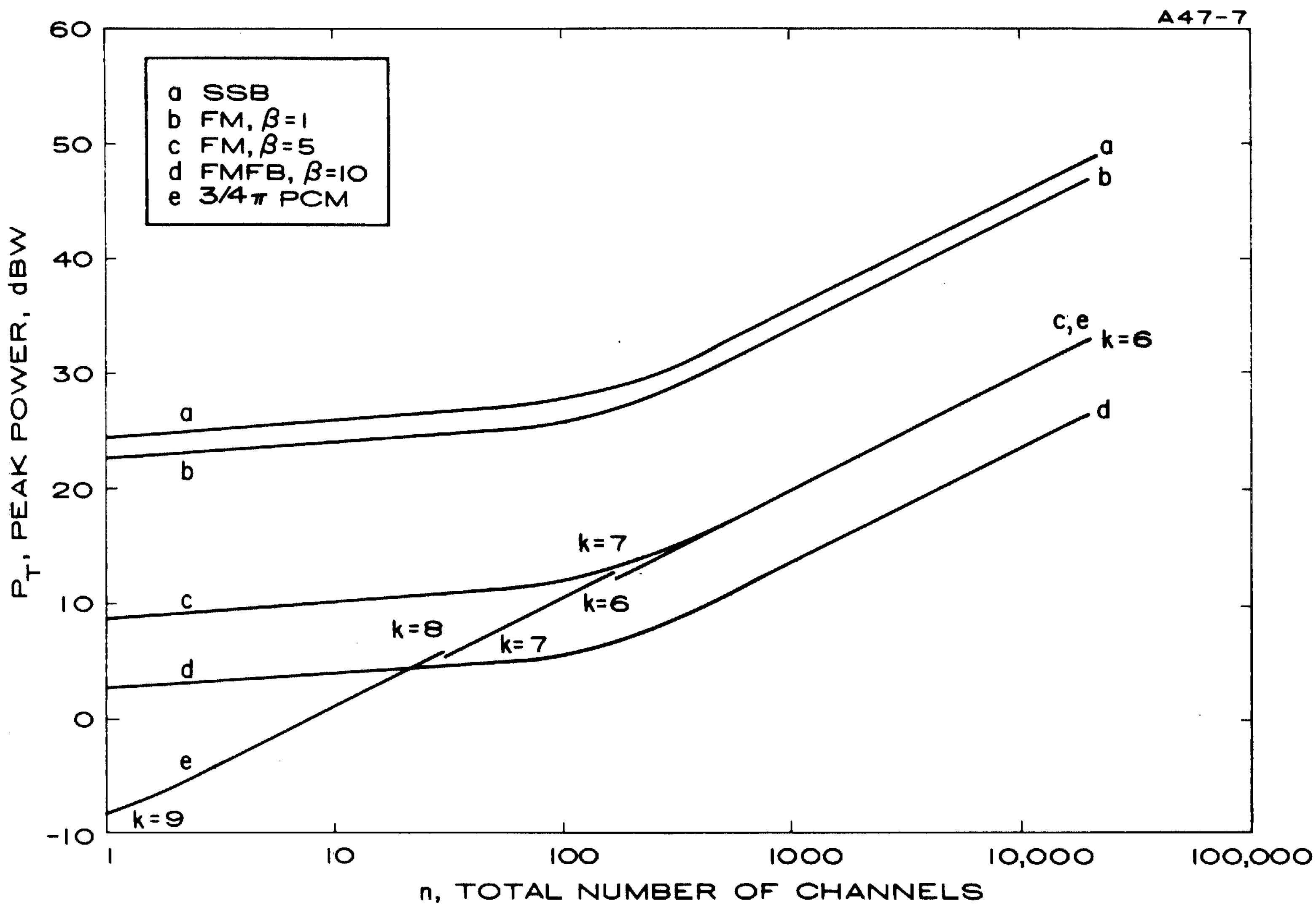

Fig. 5. Peak transmitter power, 1.0% time overload.

$$P_{T_{SSB}} = 136.2 + 44 - 15 + 10 \log n + 10.09 + 0.673 (\log n - 3.55)^2$$

$$- 167.8, \quad n \geq 240 \text{ and } \epsilon = 0.001$$

$$= 7.49 + 10 \log n + 0.673 (\log n - 3.55)^2, \quad n \geq 240$$

$$\text{and } \epsilon = 0.001$$

$$= 136.2 + 44 - 1 + 4 \log n + 10.09 + 0.673 (\log n - 3.55)^2 -$$

$$167.8, \quad 12 < n < 240 \text{ and } \epsilon = 0.001 \qquad (8)$$

$$= 21.49 + 4 \log n + 0.673 (\log n - 3.55)^2, \quad 12 < n < 240$$

$$\text{and } \epsilon = 0.001 \qquad (9)$$

Equations (8) and (9) give curve (a) in Fig. 4, and (9) can be reasonably extended to $n = 1$ as explained previously. Similar equations can be obtained for $\epsilon = 0.01$ which give curve (a) in Fig. 5, again extending the CCIR loading to $n = 1$.

Relative to actual earth (or satellite) transmitter powers, one should remember that the assumed 300°K receiver temperature is 7 dB more than 60°K (earth receiver) and 10 dB less than 3000°K (satellite receiver). Thus, for these receiver temperatures, earth transmitters would need 10 dB greater peak power, whereas satellite transmitters would need 7 dB less. As an example, for 600 channels, the 0.1% peak channel loading factor is 23 dB. With 300°K receivers,

$$P_{T_{SSB,\ 600}} = 12.4 \text{ dBW} + 23 \text{ dB} = 35.4 \text{ dBW, or } 3.47 \text{ kW.}$$

With a 3000°K satellite receiver, the earth transmitter peak power would become 34.7 kW.. Similarly, were SSB to be used from satellite to a 60°K earth receiver (admittedly unlikely in the foreseeable future) the satellite transmitter's peak power would need to be 694 W.

B. FM Calculations

Before proceeding it is necessary to consider the multiple-access aspects of FM, particularly in the up link, ground to satellite. In order to have multiple access, a number of independent ground transmitters must be able to have their signals received by the satellite, multiplexed, and retransmitted. In the complete SSB - FDM system the multiplexing is straightforward, whereas in the FM case there are several ways that groups of channels from different transmitters can be handled by the satellite.

Insofar as the earth transmitters are concerned, each would frequency modulate its FDM channels on a carrier. Choice of their carrier frequencies presents only two possibilities; either each station would use a separate carrier frequency, sufficiently spaced from all others, or it could be attempted to synchronize all carriers to arrive at the satellite with the same frequency and all in phase. Clearly the latter attempt can be discarded as seemingly hopeless, if for no other reason than that the uncontrollable phase changes in transmission through clouds and other atmospheric irregularities. There then remains the question of how best to deal with these separate carriers at the satellite.

A first approach would be to employ a simple (?) repeater to retransmit all these modulated carriers after translation to the 4 Gc "down" frequency. One objection to this approach is that it would require a much larger satellite transmitter to transmit n channels on each of p carriers than to transmit n x p channels on a single carrier. From the viewpoint of a multiple-access system of many stations, a possibly worse objection is that each earth receiver would be forced to separate and detect all these carriers in order to select its own channels only. Other objections need no discussion.

A second approach, directed toward reducing the satellite transmitter power, would be the FM-FM approach of frequency modulating all these FM signals (carriers and their sidebands) on a new carrier. This course would increase the bandwidth of the down channel, probably prohibitively, after which it would burden all of the earth receivers with still another FM detection. Both of these approaches seem to have been considered and abandoned frequently.

The third approach consists of detecting each FM earth transmission at the satellite and recombining its channels with all others into a frequency-division multiplex, for FM retransmission on a single carrier. Insofar as the earth receivers are concerned, such a signal would offer the same ease of channel selection as if transmission to the satellite had been via SSB with FM retransmission. This approach has been studied extensively in England where many engineers consider it feasible and attractive. However, careful examination of block diagrams of such a system will disclose at least the following two shortcomings: (1) It places too heavy a signal-processing burden on the satellite, requiring a receiver and FM detector for each earth station, plus the band-pass filters, frequency translators, etc., needed to regroup the channel groups into the multiplexed baseband prior to retransmission. All of this equipment might be miniaturized to be within the weight capability of present boosters and (with enough effort and cost) its reliability might be made adequate. None the less, a simpler repeater would remain preferable. (2) This approach constrains multiple access to no more stations than the number of receivers in the satellite and may further constrain their channel capacity and randomness of intercommunication so as to be best for the "permanent-channel" method of multiple access.

Despite the above objections, this third approach seems the least objectionable way of using FM from earth to satellite in a multiple-access system. Consequently, the following FM system calculations will be directed toward applicability to such a system, for which there would be interest in transmitter power versus FM bandwidth and number of channels. Additionally, FM (or PM) holds great interest for satellite to earth use in systems employing SSB from earth to satellite. Finally, none of the preceding objections apply to two-station (nonmultiple access) use of FM.

Calculations of FM transmitter power can be made similarly to those for SSB, except for the inclusion of FM improvement factors as discussed in Appendix A. Equation 6A may be expressed in decibel form as

$$P_c = \left(\frac{S}{N}\right)_c - 20 \log \left(\sqrt{3/2}\,\beta\right) - 10 \text{ Log } 2\,(\beta + 1) + P_N \text{ dBW} \tag{10}$$

recognizing that $\beta = \Delta f/nB_{ch}$, the peak frequency deviation ratio, and using the "Carson's law" FM bandwidth, $B_{FM} = 2\,n\,B_{ch}\,(\beta + 1)$. Also, C/N the receiver input carrier to noise ratio (in dB) has been replaced by its power equivalent, P_C (dBW) - P_N (dBW) before transposing P_N. $(S/N)_c$ is the "composite peak instantaneous signal to noise ratio". This P_c is interpreted as the instantaneous peak value of the input carrier power (twice its average power), and P_N is the average input noise power so that $(S/N)_c$ is interpreted as the ratio in dB of the peak power (exceeded x% of the time) of the composite (n-channel) output signal to the average power of the output baseband noise. Alternatively, when P_c is interpreted as an average power, $(S/N)_c$ must be interpreted as the ratio in dB of the average power of a multichannel sinusoidal test tone (of which the peak equals the x% peak of the composite signal) to the baseband output noise average power. Either way, a $\sqrt{2}$ average to peak (sinusoidal) voltage conversion is involved.

The instantaneous peak carrier power required of the transmitter output in our reference system is given by

$$P_{T_{FM}} = L_t + P_c \tag{11}$$

where P_N in eq. (10) is the noise power at 300°K in B_{FM}.

$$P_N = 10 \log (KTB_{FM}) = 10 \log (KTB_{ch}) + 10 \log 2 n (\beta + 1)$$

$$= -167.8 + 10 \log n + 10 \log 2 (\beta + 1) \tag{12}$$

and P_c from (7) and (10) is given by

$$P_c = 44 + l_f + p_f - 10 \log n - 10 \log (3\beta^2/2) - \log 2 (\beta + 1) + P_N \tag{13}$$

Thus from eqs. (11), (12), and (13), in which L_t, the total transmission loss is 136.2 dB,

$$P_{T_{FM}} = (10.6 - 10 \log \beta^2 + L_{fp}) \text{ dBW} \tag{14}$$

where $L_{fp} = l_f + pf$ and can be read from Figs. 3(a) and 3(c). From (1), (2) and (14),

$$P_{T_{FM}} = 5.69 + 10 \log n + 0.673 (\log n - 3.55)^2 - 10 \log \beta^2, \quad \text{for } n \geq 240, \ \epsilon = 0.001 \tag{15}$$

$$= 19.69 + 4 \log n + 0.673 (\log n - 3.55)^2 - 10 \log \beta^2, \quad \text{for } 12 \leq n \leq 240, \ \epsilon = 0.001 \ . \tag{16}$$

As before, eq. (16) can be "reasonably" extended to $n = 1$. Similar equations can be easily obtained for $\epsilon = 0.01$.

From (14), (15), and (16) one might falsely conclude that the transmitter power could be lowered indefinitely by continued increase in β. Such power reduction, however, is limited by FM threshold.

This FM threshold point is generally agreed to be about 12 dB so that the value of β must satisfy eq. (17).

$$\left(\frac{S_{av}}{N}\right)_{ch} + L_{fp} - 10 \log 3 \beta^2 (\beta + 1) \geq 12 \qquad (17)$$

where for $n \gtrsim 400$, the peak/rms, Pb from eq. (1) is about 10.3.

$$\left(\frac{S_{av}}{N}\right)_{ch} - 15 + 10.3 - 10 \log 3 \beta_{max}^2 (\beta_{max} + 1) = 12 \qquad (18)$$

For a 44 dB channel test tone to noise ratio, (18) gives $\beta_{max} \approx 5.6$.

If feedback detection of the FM, FMFB is used (see ref. 13), the threshold is reduced as given in Appendix A. Then for a detector i.f. bandwidth of three times the baseband, the threshold is reduced as given by (19).

$$\left(\frac{S}{N}\right)_{c} - 10 \log 3\beta^2 (\beta + 1) \geq 12 - 10 \log \frac{2}{3} (\beta + 1) \qquad (19)$$

Note that FMFB is only advantageous for large β. Again for $n \gtrsim 400$, eq. (20) results

$$\left(\frac{S_{av}}{N}\right)_{ch} - 15 + 10.3 - 10 \log 3 \beta_{max}^2 (\beta_{max} + 1) = 12 - 10 \log \frac{2}{3} (\beta_{max} + 1). \qquad (20)$$

Solving (20) for β_{max} letting

$$\left(\frac{S_{av}}{N}\right)_{ch} = 44 \text{ ab}$$

yields a β_{max} of 10.9 or a 9 dB reduction of the threshold. From experimental results a smaller dB reduction is more realistic or a $\beta_{max} = 10$.

For our purposes a maximum β of 5 and 10 will be used for FM and FMFB respectively, keeping in mind that these can be increased only as $(S/N)_{ch}$ is increased.

Equations (14) or (15) and (16) are plotted in Figs. 4(b) (c) (d) and 5(b) (c) (d).

C. PCM Calculations

The use of PCM with FDM means that n FDM-SSB channels are converted to PCM to form an n channel subgroup and then p subgroups are transmitted as p FDM-PCM groups. Our concern here is the peak transmitter power required for the n channel PCM subgroup as well as the random multiple-access restrictions. Regarding the latter, because PCM is essentially wideband transmission similar to FM, multiple-access possibilities are identical: (1) transmission of p subgroups FDM using a common carrier which requires phase synchronization of the carrier at the satellite receiver, (2) transmission of the p subgroups using separate carriers and simple retransmission from the satellite which requires considerable extra signal processing at each group receiver as well as additional power from the satellite transmitter, and (3) transmission of the p subgroups using separate carriers but detecting each subgroup in the satellite and retransmitting the np channels on one carrier. As with FM, method (3) seems to be the best even though the satellite complexity would be increased considerably and there would be a reduction in multiple-access flexibility with all three methods. An important aspect of multiple access is to allocate channels to each ground station according to demand which varies throughout each day. While PCM and FM can theoretically provide this service, the technical aspects are much more complicated than with SSB (see ref. 6). The major difficulty lies with varying the number and frequency of the n channels within each subgroup and programming the satellite detection processing. In spite of these multiple-access difficulties, PCM is still a possible modulation system and the power requirements remain to be determined.

PCM like FM is a modulation process which trades bandwidth for transmitted signal power. Errors in detection are directly related to the received (S/N) ratio and result in fluctuation or thermal-like noise in the output. Another source of noise is quantization noise which is solely related to the number of quantization levels or bits per sample in a binary code. This noise is present in the output regardless of the transmitted (S/N). In the end, both noise powers must be added up and the resulting output S/N is less than either S/N taken separately. However, since a substantial reduction in output fluctuation noise (lower error rate, P_e) can be obtained by small increases in transmitted signal power, it seems appropriate to specify a received S/N such that the only output noise of the system will essentially be due to quantization. It is therefore necessary to determine the received S/N for a prescribed P_e. In ref. 14 it is noted that for $P_e \approx 10^{-5}$, received S/N of 10 dB is required. It is also noted parenthetically that there is little difference in performance between binary coherent detection and binary phase comparison detection as long as $P_e < 10^{-3}$, the difference in transmitted S/N being < 1 dB. Actually, from ref. 14, experimental data indicates a larger value of S/N than the theoretical calculations. Therefore, a value of 12 dB will be assumed here.

As a matter of completeness, it is pointed out that a different approach could be used. In ref. 15 a solution for output S/N as a function of received S/N is given. This accounts for the fluctuation noise in the output caused by detection errors and neglects quantization noise. If we now decided to design the output S/N for fluctuation noise 10 dB higher than that for quantization noise, the latter would be the only essential noise in the output. Since we are using π phase modulation and ref. 15 considered only on-off pulses, a correction must be used. Several authors have included a factor of two for this correction; however, their rationale is not clear to this author. At any rate, including the factor of two and proceeding, it is noted that the required received S/N is reasonably close to the 12 dB figure assumed above in the output S/N range of interest.

One further point is that this 12 dB will be used here to specify the peak carrier to noise ratio of the PCM system. This assumption appears justified because of the difference between experimental and theoretical results as well as the fact that the FM systems to be used for comparison will inherently contain a certain small but unspecified amount of additional noise because of distortion from bandlimiting the RF spectrum.

PCM systems, in general, require transmission bandwidths twice the minimum theoretical value. However, it appears that vestigial sideband techniques (see ref. 7) can be used which require only 3/2 minimum theoretical bandwidth and that is the value which will be considered here. Also, since π phase modulation requires 2 times minimum bandwidth, the system will be designated as $3/4\ \pi$ PCM and uses k bits per sample. Equations (21) - (24) follow similarly to those for SSB and FM

$$P_{T_{(3/4)\pi\,PCM}} = \left[\left(\frac{C}{N}\right)_{(3/4)\pi\,PCM} + A + L - G_R - G_S + P_N \right] \text{dB} \tag{21}$$

$$\left(\frac{C}{N}\right)_{(3/4)\pi\,PCM} = 12 \text{ dB} \tag{22}$$

$$B_{(3/4)\,\pi\ PCM} = (3/2)\, n\, k\, B_{ch} \text{ cps} \qquad (23)^*$$

$$P_N = (-167.8 + 10 \log (3/2)\, nk) = -166 + 10 \log nk \text{ dBW} . \quad (24)$$

The value of k must be determined from the $(S/N)_c$ for the n channels given by (7) and (25).

$$\left(\frac{S}{N}\right)_c = \left(\frac{S_{av}}{N}\right)_{ch} + l_f + p_f - 10 \log n \approx 6\, k + 4.8 \text{ dB} . \qquad (25)^*$$

For a $(S_{av}/N)_{ch}$ of 44 dB in this case and using (1) and (2),

$$6\, k + 4.8 \approx 44 + (-15 + 10 \log n) + 10.09 + 0.673 (\log n - 3.55)^2 - 10 \log n,$$

$$\approx \quad \text{for } n > 240, \ \epsilon = 0.001 \qquad (26)$$

$$44 + (-1 + 4 \log n + 10.09 + 0.673 (\log n - 3.55)^2 - 10 \log n,$$

$$\text{for} \approx 12 < n < 240, \ \epsilon = 0.001 \qquad (27)$$

Solving (26) and (27) for k remembering that k can really only assume integer values gives

$$k \approx 5.71 + 0.112 (\log n - 3.55)^2 \approx 6.0 \quad , \quad 240 < n < 10^4, \ \epsilon = 0.001 \qquad (28)$$

$$\approx 8.05 - \log n + 0.112 (\log n - 3.55)^2 , \quad 12 < n < 240, \ \epsilon = 0.001 \qquad (29)$$

so that for $\epsilon = 0.001$,

*PCM bandwidth and S/N ratio is discussed in more detail in Section V.

$$
\begin{aligned}
k &= 8,\ 12 < n < 35 \\
&= 7,\ 35 < n < 170 \qquad (30) \\
&= 6\ n > 170
\end{aligned}
$$

$$
\begin{aligned}
\text{and } (S/N)_c &= 53 \text{ dB},\ 12 < n < 35 \\
&= 47 \text{ dB},\ 35 < n < 170 \qquad (31) \\
&= 41 \text{ dB},\ n > 170 \quad .
\end{aligned}
$$

The actual output of the PCM system will have somewhat better S/N ratio than that specified because of the smaller quantization noise here than thermal noise in the analogue systems.

From the above values of k, (21)-(24), and (5)

$$
\begin{aligned}
P_{T_{(3/4)\pi PCM}} &= (12 + 136.2 - 166 + 10 \log n + 10 \log k) \text{ dBW} \qquad (32) \\
&= (-17.8 + 10 \log n + 10 \log k) \text{ dBW} \\
&= (-8.8 + 10 \log n) \text{ dBW},\ 12 < n < 35 \qquad (33) \\
&= (-9.3 + 10 \log n) \text{ dBW},\ 35 < n \text{ S } 170 \qquad (34) \\
&= (-10 + 10 \log n) \text{ dBW},\ n > 170 \quad . \qquad (35)
\end{aligned}
$$

For one channel, as before, there is - 1 dB channel loading and 18.4 dB peak/rms loading. Solving for k in (29) gives $k \approx 10$. Therefore

$$
P_{T_{(3/4)\pi PCM}} = -7.8 \text{ dBW for one channel.} \qquad (36)
$$

A curve of (33), (34), (35), and (36) is shown in Fig. 4. An identical approach for $\epsilon = 0.01$ has been carried out and the results plotted in Fig. 5.

V. GENERALIZED SYSTEM COMPARISON

A. Over-all Development

The previous section compared the three modulation systems, SSB, FM, and PCM, on the basis of a particular reference system and varying number of channels. Also, the prime quantity for comparison was the peak instantaneous transmitter power because the cost of a transmitter will depend primarily on the peak power rather than on the rms or average power. Bandwidth is generally only of secondary importance as far as cost is concerned. What is desired now is a more generalized comparison which is independent of the reference system parameters. There are a number of possible approaches to do this but one which seems most appropriate was the comparison from an "information efficiency" standpoint where quotation marks are used because a modified rate comparison is used rather than actual bit rates. Perhaps "information measure" or "system utilization" might have been better terminology so the reader is cautioned to keep in mind the way the term "information efficiency" is used in this development. The desired results are a comparison of peak power required for each system as a function of bandwidth and composite S/N ratio.

For this comparison we begin with Shannon's ideal or maximum theoretical bit rate per channel, I_{ch}.

$$I_{ch} \simeq B_{ch} \log_2 \left[1 + \left(\frac{S_{av}}{N} \right)_{ch} \right] \text{bits/sec.} \tag{37}$$

where

B_{ch} = information bandwidth or channel bandwidth

$1 + 2B_{ch} = 2B_{ch}$ = samples/sec required from the Sampling Theorem, (see ref. 16)

$\sqrt{\left(\frac{S_{av}}{N} \right)_{ch} + 1}$ = average number of distinguishable voltage levels as seen in Fig. 6.

$\left(\frac{S_{av}}{N} \right)_{ch} = \frac{\text{average signal power per channel}}{\text{mean square thermal noise power in } B_{ch}}$

A one is added to account for the zero level; distinguishable voltage levels are the square root of the distinguishable power levels; and the approximation sign in (37) is used because the equation is valid on the average only. This

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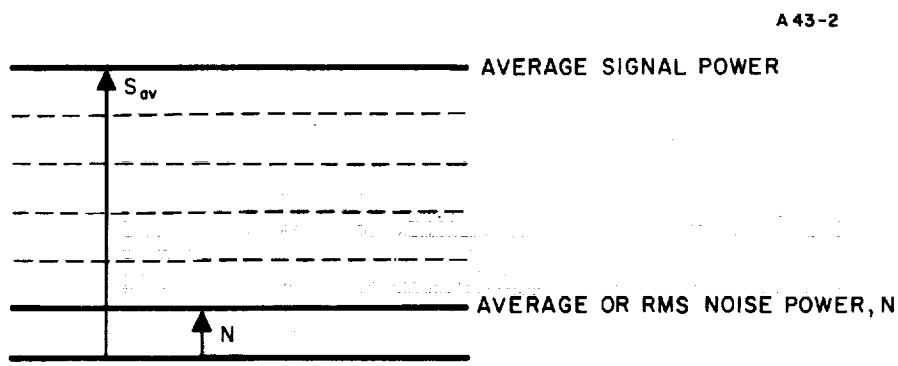

Fig. 6. Power level diagram.

bit rate is the theoretical information rate possible from a SSB System. If such a system were converted to an equivalent binary digital pulse system there would be k bits/sample specifying 2^k voltage levels; thus

$$2^k \approx \sqrt{1 + \left(\frac{S_{av}}{N}\right)_{ch}} \text{ levels} \tag{38}$$

$$k \approx \frac{1}{2} \log_2 \left[1 + \left(\frac{S_{av}}{N}\right)_{ch}\right] \text{bits/sample} \tag{39}$$

and the maximum information rate of the equivalent binary system would be $2kB_{ch}$ which results in (37). Therefore, considering the average signal power only, one can conclude that SSB is 100 % efficient from an information standpoint.

This heuristic argument is given only to provide a certain plausibility to the final general comparison to be used. It is noted that if the signal is noise-like, i.e., its probability density is gaussian, then the $(S_{av}/N)_{ch}$ is equal to $(S/N)_{peak}$ (ratio of peak signal power to peak noise power). This would be true for the $(S/N)_c$ when $n > 240$ but would not be true for $n < 240$. Therefore, the expression for information as applied to this generalized comparison of modulation systems is modified as given by (40).

$$I_{SSB} = nB_{ch} \log_2 \left(\frac{S}{N}\right)_c \tag{40}$$

where the signal power is the peak instantaneous power, the average noise power in nB_{ch} is used, and the one has been dropped because it adds nothing to the comparison desired $(S/N)_c >> 1$ anyway. In other words, the previous heuristic argument could be deleted and the development begun with eq. (40) by stating that a comparison will be made by weighting bandwidth directly and (S/N) logarithmically. This I_{SSB} will be used as the comparison reference for expressing information efficiencies of the other modulation systems in percent relative to SSB.

Regarding FM and PCM we find that the actual or equivalent theoretical maximum information bit rate in the output, I_{ch}, is considerably less than the maximum theoretical rate as specified by (37) when the RF bandwidth and ratio of average carrier power to RF noise power are used. Since "information" as used here is similar to information bit rate, information efficiency is defined as the ratio of actual "information", I_{SSB}, to theoretical "information".

For FM then, the information efficiency (or measure of system utilization) including the modifications of eq. (40) is given by

$$\mathrm{Eff}_{FM} = \frac{nB_{ch} \log_2 \left(\frac{S}{N}\right)_c}{B_{FM} \log_2 \left(\frac{C}{N}\right) FM} \tag{41}$$

where B_{FM} and $(C/N)_{FM}$ are given in Appendix A.

If Eff_{FM} is denoted by x, the ratio of peak transmitter power to average noise in the RF bandwidth can be obtained for either system knowing the value for the other from (41).

$$\left(\frac{S}{N}\right)_c = \left(\frac{S}{N}\right)_{SSB} \mathrm{dB} = \left(\frac{C}{N}\right)_{FM} \frac{x^{B_{FM}}}{B_{SSB}} \mathrm{dB} \tag{42}$$

where $(B_{FM}/B_{SSB} = 2\ (\beta + 1)$ and $(C/N)_{FM} = (1/3\beta^2\ (\beta + 1)\ (S/N)_c$ from Appendix A. Thus

$$\mathrm{Eff}_{FM} = \frac{\log_2 \left(\frac{S}{N}\right)_c}{2(\beta+1) \log_2 \left[\frac{1}{3\ \beta^2\ (\beta+1)} \left(\frac{S}{N}\right)_c\right]} =$$

$$\frac{\left(\frac{S}{N}\right)_c \ ab}{2\ (\beta+1) \left\{\left(\frac{S}{N}\right)_c \ ab - \left[3\beta^2\ (\beta+1)\right] \ ab\right\}} \tag{43}$$

Equation (43) is plotted in Fig. 7 for several values of β.

Now we turn our attention to PCM where it is noted that the $(S/N)_c$ can only assume certain discrete values because k must be an integer in any practical system. The $(C/N)_{PCM}$ was given previously as 12 dB so that output noise will only be N_q, that which is caused by quantization. Concern is, therefore, directed toward (S/N_q) and RF bandwidth for PCM. From ref. 16 as well as other sources, the <u>rms signal power</u> to quantization noise is given by

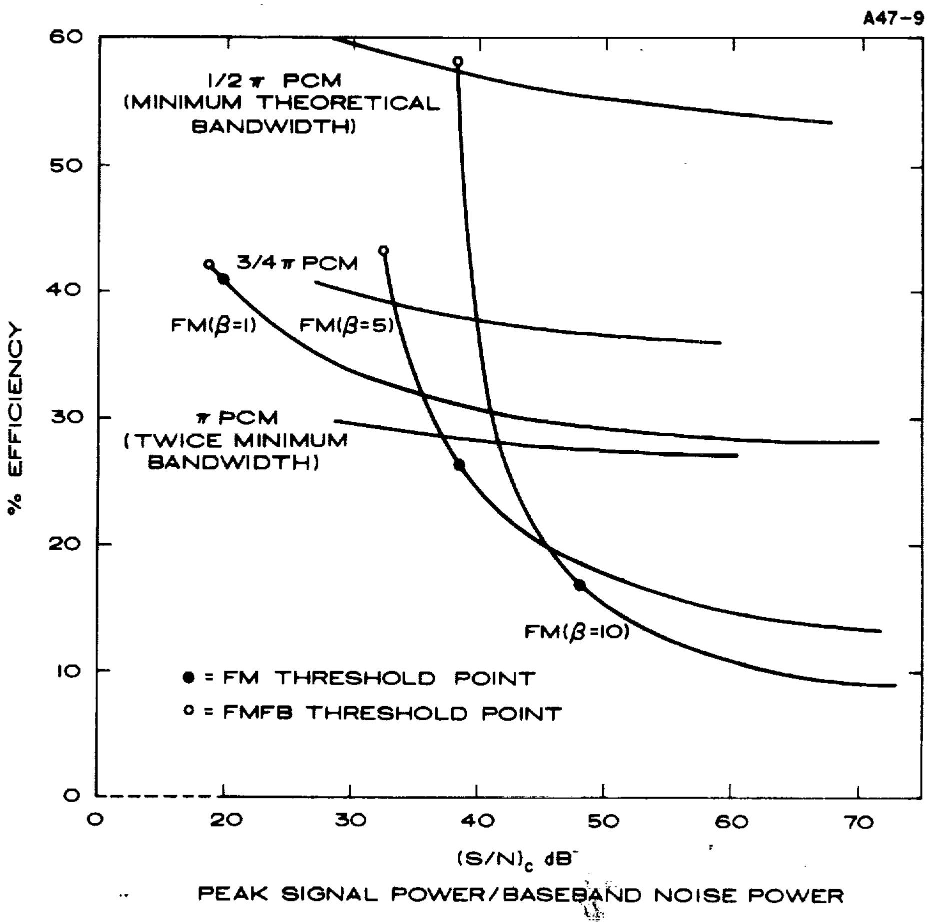

Fig. 7. Information efficiency relative to ideal SSB.

$$\frac{S}{N_q} \text{ dB} = (1.76 + 6k) \text{ dB} \tag{44}$$

$$\frac{S}{N_q} = 3/2 \times 2^{2k} \tag{45}$$

$$\left(1 + \frac{S}{N_q}\right) = 2^{2k}\left[2^{-2k} + 1.5\right] \tag{46}$$

$$\log_2\left(1 + \frac{S}{N_q}\right) \simeq 2k + 0.585, \text{ for large k} . \tag{47}$$

Since there are ideally 2k bits per second per cycle of baseband frequency,

$$\begin{aligned} 2knB_{ch} &= nB_{ch}\left[\log_2\left(1 + \frac{S}{N_q}\right) - 0.585\right], \quad \text{for large k} \\ &\simeq nB_{ch} \log_2\left(1 + \frac{S}{N_q}\right) \\ &\simeq nB_{ch} \log_2\left(\frac{S}{N_q}\right), \quad \text{for large } (S/N_q) \text{ and} \\ &\qquad \text{large k.} \end{aligned} \tag{48}$$

In fact, had the actual rms power in the quantized sine wave signal been used rather than the rms of a sine wave, the answer would be exact rather than approximate. Since (48) is identical to Shannon's ideal rate, we have shown that quantization noise in PCM is identical to thermal noise in analogue systems, as anticipated, at least with respect to information. For our purposes (44) must be modified for peak instantaneous signal power as given by (49).

$$\left(\frac{S}{N}\right)_c = (6\text{ k} + 4.8) \text{ dB} \tag{49}$$

With regard to PCM bandwidth, it is theoretically possible to extract 1 bit of information for 1/2 cps of bandwidth. However, in practical systems like π phase modulation 1 cps of bandwidth is required for each bit so that

$$B_{\pi PCM} = 2\,k\,n\,B_{ch} \quad . \tag{50}$$

This same bandwidth is required of double sideband AM pulse transmission. If single sideband transmission of on-off pulses can be developed, there is a reduction of the bandwidth by 2.

$$B_{(1/2)\pi PCM} = k\,n\,B_{ch} \quad . \tag{51}$$

Both (50) and (51) will be used here with the appropriate subscript in order to show the difference between a normal practical system and that theoretically possible. Previous calculations in Section IV using the reference system assumed 3/2 minimum bandwidth or 3/4 the π phase system bandwidth as being the smallest bandwidth believed to be practical.

Information efficiency is expressed in modified form as compared to SSB in similar manner to the FM case above. For π phase modulated PCM systems, (50),

$$\mathrm{Eff}_{\pi PCM} = \frac{nBch \log 2\left(\frac{S}{N}\right)_c}{B_{\pi PCM} \log_2\left(\frac{C}{N}\right)_{\pi PCM}} = \frac{(6k + 4.8)\ \mathrm{dB}}{2k\ (12\ \mathrm{dB})} = 0.25 + \frac{0.2}{k} \tag{52}$$

where twice the theoretically minimum RF bandwidth is assumed and the $(S/N)_c$ is taken as (6 k + 4.8) which is the peak signal to quantization noise.

$$\mathrm{Eff}_{(3/4)\pi PCM} = 0.33 + \frac{0.267}{k} \tag{53}$$

and for PCM systems using ideal minimum RF bandwidth the information efficiency is simply twice as large as $\mathrm{Eff}_{\pi PCM}$

$$\mathrm{Eff}_{(1/2)\pi PCM} = 0.5 + \frac{0.4}{k} \quad . \tag{54}$$

Thus PCM appears to trade bandwidth for (S/N) ratio very nearly as an ideal system would. The difficulty, of course, is that one cannot trade smaller (C/N) ratio for larger RF bandwidths and vice versa. Equations (52), (53) and (54) are plotted in Fig. 7.

An expression similar to (42) can also be written for comparison of SSB with PCM, but $(S/N)_c$ can only have discrete values given by (6 k + 4.8).

B. Comparison of FM and PCM

Perhaps the most meaningful results from Fig. 7 is the comparison of FM and PCM systems. For this comparison let $Eff_{\pi PCM} = y$. Then the $\log_2 (S/N)_c$ will cancel when a ratio of efficiencies is taken providing the same value is used in both systems.

$$\frac{y}{x} = \frac{2(\beta+1)\log_2\left[\frac{C}{N}_{FM}\right]}{2\,k\,\log_2\left[\frac{C}{N}_{\pi PCM}\right]} = \frac{(\beta+1)}{k}\,\frac{\left(\frac{C}{N}\right)_{FM}\,dB}{\left(\frac{C}{N}\right)_{\pi PCM}\,dB} \quad . \tag{55}$$

Solving for $(C/N)_{FM}$ and using 12 dB for $(C/N)_{\pi PCM}$,

$$\left(\frac{C}{N}\right)_{FM} = \frac{y}{x} \times \frac{k}{(\beta+1)} \times 12 \text{ dB} \quad . \tag{56}$$

If one wishes to calculate the peak carrier to average noise when a different $(S/N)_c$ is used in each system, the results are

$$\left(\frac{C}{N}\right)_{FM} = \frac{y}{x} \times \frac{k}{(\beta+1)} \times 12 \times \frac{\left(\frac{S}{N}\right)_{SSB}}{6\,k + 4.8} \tag{57a}$$

$$= \frac{y}{x} \times \frac{B_{PCM}}{B_{FM}} \times 12 \times \frac{\left(\frac{S}{N}\right)_c}{\left(\frac{S_c}{N_q}\right)} \tag{57b}$$

where $(S/N)_{SSB}$ is the $(S/N)_c$ at the output of the FM system.

These comparisons, (55) and (57), assume that different RF bandwidths are used for each system, so now the comparison can be made fixing β to yield a $B_{FM} = B_{PCM}$. Additionally, we will now consider the $(3/4)_{\pi}PCM$ system because of its practical application. A comparison of bandwidths is given in Table III.

TABLE III

Relative System Bandwidths

System	Bandwidth
SSB	1
FM	$2(\beta + 1)$
$(3/4)\pi$PCM	$(3/2)k$

Thus for equal FM and PCM bandwidths,

$$\beta = 0.75k - 1 . \tag{58}$$

The comparison of these systems is given in Table IV, of which the data are plotted in Fig. 8, drawing a smooth curve through the discrete points from eqs. (43) and (53).

TABLE IV

Comparison of $(3/4)\pi$PCM with FM for Equal Bandwidths

$(S/N)_c$	23	29	35	41	47	53
k	3	4	5	6	7	8
β	1.25	2.0	2.75	3.5	4.25	5.0
$Eff_{\frac{3}{4}\pi PCM}$	0.42	0.40	0.387	0.378	0.371	0.366
Eff_{FM}	0.50	0.36	0.297	0.242	0.20	0.167

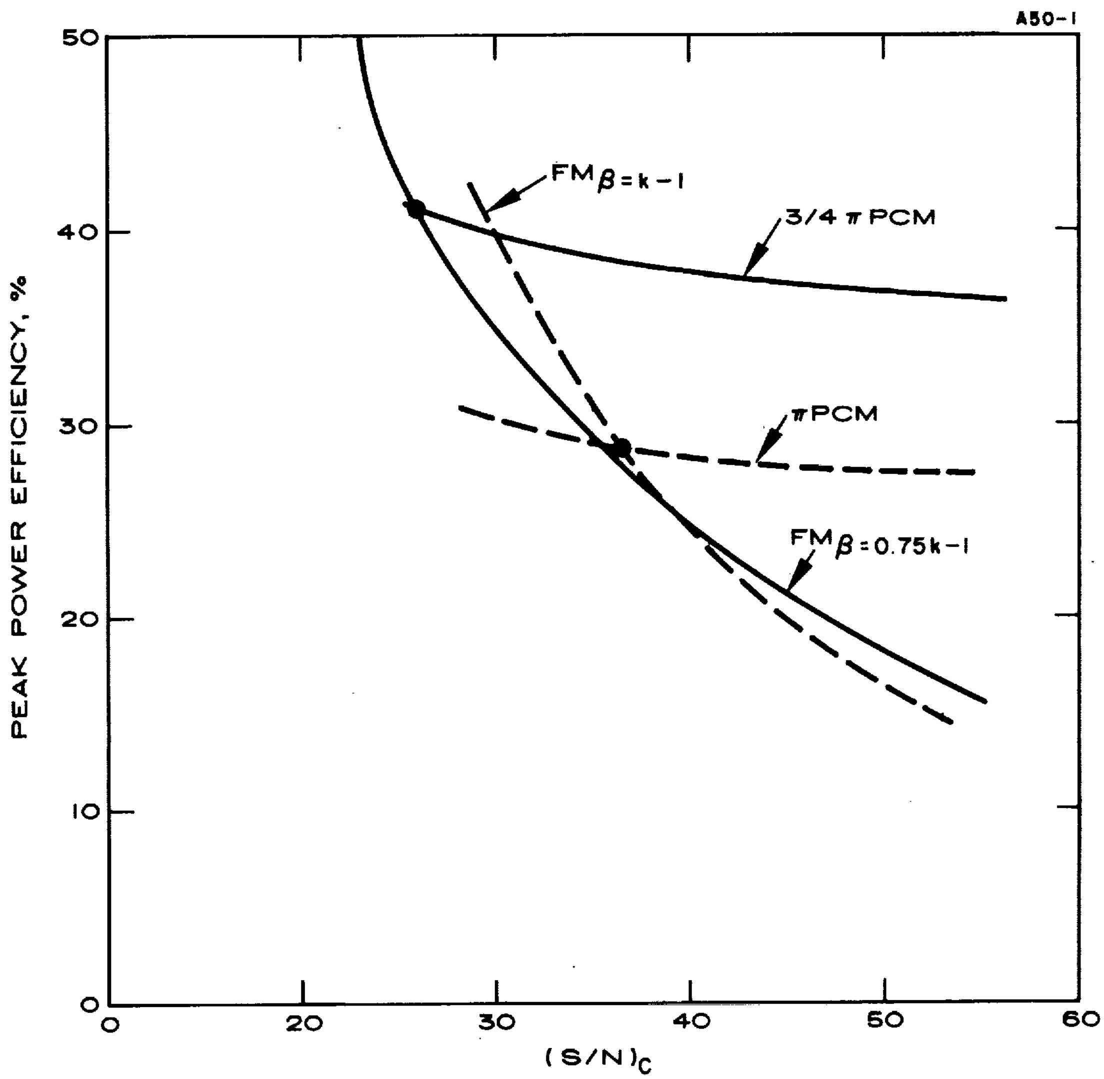

Fig. 8. PCM and FM efficiencies for equal bandwidth systems.

A similar calculation for πPCM using (43) and (52) is given in Table V for which $\beta = k - 1$. This data is given in Fig. 8 by the dashed curves.

TABLE V

Comparison of πPCM with FM for Equal Bandwidths

$(S/N)_c$	29	35	41	47	53
k	4	5	6	7	8
β	3	4	5	6	7
$Eff_{\pi PCM}$	0.30	0.29	0.283	0.279	0.275
Eff_{FM}	0.418	0.312	0.236	0.184	0.149

The significant result from Fig. 8 is the fact that transmitter power requirements of FM systems as compared with PCM systems are a function of $(S/N)_c$. For large (S/N) ratios, PCM systems require less power for the same bandwidth whereas for small (S/N) ratios, FM systems will require less power. This explains fundamentally the diverse conclusions found in the literature regarding which system is most conserving of transmitter power. Specifically, if twice minimum bandwidth is used for PCM, FM will require less power for instantaneous peak signal to average noise power, $(S/N)_c$, less than about 37 dB. When 3/2 minimum bandwidth is used for PCM, the crossover point is about 25 dB.

C. Use of the Generalized Comparison

It is possible to determine the P_T for system "a", P_{T_a}, knowing P_T for system "b", P_{T_b}, as well as the modulation system parameters. This avoids calculating back through the transmission loss and noise power each time a comparison is made.

$$P_{T_a} \simeq P_{T_b} + \left(\frac{C}{N}\right)_b \left[\frac{Eff_b}{Eff_a} \times \frac{B_b}{B_a} - 1\right] + 10 \log \frac{B_a}{B_b} \quad (59)$$

where

$(S/N)_c \simeq 6k + 4.8$ is assumed

B_a = RF bandwidth of system a

B_b = RF bandwidth of system b

$(C/N)_b$ = peak carrier/average noise in system b = $(S/N)_c \big|$ b = SSB.

The following illustrative examples are given to show how the results might be applied:

1. Use of eq. (57). Assume a 1000 channel system, $(S/N)_{ch}$ = 44 dB, ϵ = 0.001 and β = 5. From eq. (7) the $(S/N)_c$ = 39.2 dB; Eff_{FM} = 0.256 from Fig. 7 [or (43)]; and B_{FM} = 12 nB_{ch}. For (3/4) πPCM, k = 6, Eff_{PCM} = 0.378 from Fig. 7 [or (53)]; 6 k + 4.8 = 40.8 dB; and B_{PCM} = 9 nB_{ch}. Since $(C/N)_{PCM}$ = 12 dB, from (57)

$$\left(\frac{C}{N}\right)_{FM} = \frac{0.378}{0.256} \times \frac{9}{12} \times \frac{39.2}{40.8} = 12.7 \text{ dB} .$$

This example shows that (57) by itself is not very useful because $(C/N)_{FM}$ could have been calculated directly with greater ease. Without knowing $P_{T_{PCM}}$, the total transmission loss and noise power must be used to obtain $P_{T_{FM}}$.

2. Use of eq. (59). Assume n = 12; $(S/N)_{ch}$ = 44 dB; ϵ = 0.001, and β = 5. Let $P_{T(3/4)\pi PCM}$ = 2 dBW (from Fig. 4e). From eq. (7) the $(S/N)_c$ = 50.7 dB which is very nearly equal to 6 k + 4.8 since k = 8 which gives (S_c/N_q) = 52.8. From Fig. 7 $Eff_{FM(\beta = 5)}$ = 0.175 and $Eff_{3/4\ \pi\ PCM}$ = 0.365. From (57)

$$\left(\frac{C}{N}\right)_{FM} = \frac{0.365}{0.175} \times 12 \times \frac{50.7}{52.8} = 24 \text{ dB} .$$

Equation (59) can be written for this example in the form

$$P_{T_{FM}} = \left[P_{T_{PCM}} + \left(\frac{C}{N}\right)_{FM} - 12 + 10 \log \frac{B_{FM}}{B_{PCM}} \right] \text{dBW} \tag{60}$$

$$= 2 + 24 - 12 + 0 = 14 \text{ dBW}$$

This agrees with the value obtained from Fig. 4c as anticipated. Had (59) been used directly, the (50.1/52.8) term is neglected and $P_{T_{FM}}$ = 14.9 dBW results.

3. Further use of eq. (59). Use the same system as specified in example 2 above except that $P_{T_{SSB}}$ = 30 dBW is given (from Fig. 4a). From (59)

$$P_{T_{FM}} = 30 + 50.7 \left[\frac{1}{0.175} \times \frac{1}{12} - 1 \right] + 10 \log 2 (5 + 1) = 14.2 \text{ dBW} .$$

The major inconvenience in these examples is the determination of $(S/N)_c$ each time; however, it is somewhat convenient to be able to determine the P_T for two of the systems knowing the transmitter power for the third without having to include total transmission loss and noise power each time.

VI. APPLICATION TO TV

Multiple-access arguments do not apply for TV because ground transmission will be from only one transmitter in early satellite communication systems. Thus, attention is directed to single transmitter requirements or what has previously only been primarily down link considerations.

One usual specification is that the satellite communication system must be able to transmit one TV channel instead of 1200 telephone channels. Reference 5 gives the CCIR recommendation for 625 line, 5 Mc, TV systems of 52 dB weighted $(S/N)_{TV}$ ratio where

$$(S/N)_{TV} = \frac{\text{peak-to-peak video signal power}}{\text{rms noise power in total frequency band}} \tag{61}$$

Information about the weighting network is given in CCIR recommendation No. 267 where the dB weighting values of 8.5 dB for "white" noise and 16.3 dB for "triangular" noise is given. Also (61) considers peak-to-peak signal power because the TV video signal is of one polarity as compared to speech which is composed of double polarity waves. However, in previous power calculations for the speech channels, a 10.2 dB average peak to rms factor was included. Therefore, it appears appropriate to compare a 1200 channel speech system calculated previously with the TV system specifications directly without changing the signal power. Thus a

$$\begin{aligned} (S/N)_{TV} &= 52 - 8.5 = 43.5 \text{ dB unweighted for flat noise} \\ &= 52 - 16.3 = 35.7 \text{ dB unweighted for triangular noise} \end{aligned} \tag{62}$$

is required, whereas previous calculations were for a S/N ratio in the composite 1200 speech channel system of 39.2 dB. The result then is an increase of 4.5 dB in transmitted power in SSB transmission, (Fig. 4), to accommodate one TV channel within present CCIR specifications for ground microwave relay systems. FM power requirements are 3.5 dB higher in Fig. 4 than required because the noise is triangular; PCM, on the other hand, requires 2.7 dB increase in power.

It is pointed out that CCIR recommendations for tropospheric scatter TV transmission have not been made as yet. According to ref. 5 a smaller value of $(S/N)_{TV}$ is anticipated because all CCIR recommendations represent compromises between quality and cost. In the light of severe increases in cost, lower quality performance is generally considered acceptable.

Another point is that recommendations for a 4 Mc, 525 line system have not been made whereas the recommended $(S/N)_{TV}$ for a 3 Mc, 405 line system is only 50 dB weighted. Thus, one might expect that a smaller value of $(S/N)_{TV}$ would be acceptable for a TV system based on present USA standards.

Recommendations sponsored by the USA argue that the above CCIR requirements are too low. The U. S. proposal is to increase the $(S/N)_{TV}$ unweighted to 48 dB for "flat" noise and 43 dB for "triangular" noise. This, of course, would increase the power requirements of SSB appreciably as well as having smaller but increasing effects on FM and PCM power transmitted.

VII. MISCELLANEOUS ASPECTS

This report has considered the fact that power amplifiers presently available for 4 Gc and 6 Gc are all peak power limited rather than average power limited; therefore, our concern has been the peak power requirements for each modulation system. With SSB, there is the problem of providing large peak powers with accompanying gain stability and linearity. Possible future developments include depressed collector TWT power amplifiers which might be average power limited while still capable of supplying sufficient additional power for peak amplitudes in the SSB case. This would substantially alleviate the large power requirements for SSB. However, until such developmental efforts come to fruition, use of SSB requires large transmitters to supply the high peak power. An associated problem arises from the allowable total of intermodulation and thermal noise. The latter can be reduced by increased power. Given a particular power amplifier, the output power increases with the drive power but so does the intermodulation distortion. Calculation of the optimum peak power is beyond the scope of this report and allocations of thermal noise and intermodulation distortion were made arbitrarily. Also, it has been assumed that distortion specifications could be met by $0.01 \leq \epsilon \leq 0.001$. It is also noted that peak powers required for SSB would be prohibitive (for CCIR quality) without the use of compandors and that 13.5 dB companding improvement was assumed, whereas 17 or 18 dB improvements are quoted as presently available for 2:1 syllabic compandors. As has been mentioned elsewhere (see ref. 17), the further development of compandors could substantially alleviate the peak power requirements. Another development might take the form of more advanced techniques for linearization of power amplifiers as in ref. 18. Since multiple-access requirements dictate a preference for SSB in the up link, its use by many stations depends upon the economic feasibility of light-traffic stations. In turn, this feasibility depends on controlling the fixed cost of these stations. The cost of present SSB transmitters is determined primarily by the peak power requirement, and, therefore, is relatively constant for $n < 150$, from Fig. 4. Reduction of transmitter costs appears to depend upon useful compandor improvement and success of depressed collector traveling-wave tubes (or similar techniques).

With FM for the up link the multiple-access problems are severe, if not almost prohibitive. FM for the down link, on the other hand, is very attractive except for the problems created by use of large bandwidths. One of these is the development problem of practical feedback detectors for 5 Mc or larger baseband frequencies. As this report is being written the largest reported baseband capability of feedback detectors is about 3.5 or 4.0 Mc. Even without feedback detection, the satellite power requirements are not too severe so that FM presents a good solution for the down link. This is particularly true when one considers the ease of converting SSB to FM in the satellite.

As for PCM, with frequency-division multiplexing there are multiple-access problems similar to those with FM, making PCM very unattractive for the up link. Additionally, the conversion from SSB to PCM in the satellite is very complex as opposed to the easy SSB to FM conversion. Thus, as a consequence, FDM-PCM is the least attractive system because it would have to be used in both the up and down link. Note that this might not be true if PCM could be used with time-division multiplexing.

In one respect, the information efficiencies and power calculations for PCM are slightly unfair because this system is being compared with FM systems which inherently contain 3 or 4% distortion in the output at peak deviation.* A PCM system for $k = 7$ will contain about 1% distortion due to quantization noise and if distortion is treated the same as thermal noise, an additional 2% could be allowed in PCM for noise due to error rate. From ref. 19, an error rate of 10^{-3} yields approximately 1% additional noise and 10^{-2} error rate yields about 8% additional noise. Therefore, an error rate for the PCM system could have been about 10^{-3} rather than 10^{-5} as originally used. Then from ref. 14, a reduction in PCM power of approximately 3 dB could be argued in Figs. 4 and 5. A compensating effect takes place, however, when one deviates from πPCM in an attempt to reduce the RF bandwidth. This is caused by an uncertain amount of peaking in the RF carrier from deletion of sidebands.

PCM, in general, appears to present the most formidable development problems. First, and least consequential, is the percent quantization noise increasing with low amplitude signals. It would appear that a satisfactory solution to this problem can be obtained by using additional instantaneous compandors (nonlinear encoding) which would "effectively" add about 3 or 4 digits to the quantization process. The major problem appears to be the achievement of practical systems capable of, for example, 58 megabits/sec transmission rate** or perhaps even larger if more than 6 bits/sample or more than 1200 channels are used. This indeed strains the present state of the art with regard to pulse generation, nanosecond switching, and synchronization.

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*This distortion in FM appears in the upper baseband channels and merits further study which is beyond the scope of this report.

**$2 \times 6 \times 4000 \times 1200 = 57.8 \times 10^6$

Calculations for all three systems assumed the transmission of a multitude of voice telephone channels, and therefore used speech channel loading factors. Now the question arises: what kind of loading results when music or continuous radio broadcasting is transmitted? Presumably, the peak values could be limited to those allowed for speech, but it would appear that the rms level would increase. Thus, present designs would be adequate assuming, of course, that power amplifiers were peak power limited and not rms limited.

There is always the possibility of new modulation schemes arising. Perhaps one of the more interesting is single sideband frequency modulation (see ref. 20). Again, this would have satellite multiplex problems similar to FM, but the possibility of a 1/3 saving in bandwidth looks attractive. Another recent advance is given in ref. 21 where PPM is used to smooth over the quantization levels of PCM, thus requiring fewer quantization steps for the same S/N ratio and a saving in bandwidth or signal power results.

VIII. CONCLUSIONS

This study has examined the relative merits of three familiar and distinct modulation methods relative to their applicability to random multiple-access satellite communication systems. The three systems were considered the most likely to be used and if PCM should appear to be attractive, additional pulse systems should be considered. Since presently available power amplifiers for 4 Gc and 6 Gc are all peak power limited rather than average power limited, our concern has been the peak power requirements for each system.

A. Multiplexing

Common spectrum technology is relatively new and has not been reduced to common carrier practice. The tentative conclusion is that it is questionable whether the equipment and quality of service will become commercially competitive although there is the potential multiple-access advantage of direct calling.

For multiplexed analog (e.g., voice) communication, frequency division (FDM) has been most thoroughly developed and extensively used. It is generally considered to be the easiest to achieve and the most economically inexpensive. Additionally, no serious technical difficulties are anticipated for high-capacity systems. Therefore, FDM was the choice for the multiplex system.

Time-division-multiplex (TDM) is theoretically possible for multiple access. However, thus far, the application of TDM has been limited to 300 voice channels although systems of higher capacity are being developed. Additionally, for multiple-access satellite communication, TDM brings the problem of synchronizing the transmission from many stations in order for them to reach the satellite within their respective narrow time-slots. A possibly fatal aspect of TDM is its growth inflexibility, since adding channels shortens the time per channel and requires major equipment changes. Hence, it was concluded that TDM systems should be neglected in this study. It is recognized, however, that the state of the art advances and that one cannot necessarily say that TDM will remain impractical for high-capacity systems.

B. Modulation — Earth to Satellite, Up Link

Assuming the use of FDM with transmission from many earth stations, the multiplexing at the satellite is "natural" and straightforward when the various transmissions are FDM-SSB in separate channels within the satellite's receiving band. In contrast, the use of FM or PCM leads to penalizing the satellite via excessive retransmission power, excessive retransmission bandwidth and/or excessive complexity as a consequence of signal processing. Additionally, there may be loss of system flexibility and complication of the many earth receivers, as has been explained. Were it not for these apparently prohibitive troubles, either FM or PCM could be used to reduce the power (and cost) of the earth transmitters and to overcome other problems encountered with SSB, such as the high linearity of amplification necessary to control intermodulation noise. As it is, these problems attendant to the use of SSB are ones which are well understood and which all have good prospects of satisfactory solution. Hence, the fact that SSB requires high peak transmitter powers and/or the use of compandors is relatively immaterial.

Within the scope of present understanding, the use of FM or PCM in the up links of high-capacity random multiple-access systems leads to satellite multiplexing problems and to others which appear so much worse as to be prohibitive.

Neglecting these multiple-access constraints, one sees that SSB requires minimum bandwidth, whereas FM and PCM trade bandwidth for reduced power. Today, this bandwidth at satellite communication frequencies appears plentiful, so the bandwidth saving of SSB does not yet assume much importance. Neglecting multiple-access and bandwidth conservation, e.g., for TV transmission or two-terminal communication, wideband FM is most attractive from both technical and economic standpoints. However, more efficient use of bandwidth will increase in importance as the use of satellite communication increases. With FM, the deviation could be reduced to reverse its bandwidth-power trade-off, much as has been done with high-capacity microwave relay systems. However, at the lowest deviations, FM becomes less efficient than SSB because most of its power is wasted in its carrier spike, while its minimum bandwidth approaches twice that of SSB. PCM is even less flexible in its ability to trade power for bandwidth reduction because bandwidth is determined by the number of quantization levels (hence by S/N_q) and by the sampling rate. Hence, in regard to future bandwidth conservation, the present adoption of SSB would be ideal — no further bandwidth reduction need be contemplated. The adoption of PCM would be least desirable from this viewpoint because of its bandwidth inflexibility.

C. Modulation — Satellite to Earth, Down Link

For this down link the problems are distinctly different in that there is a single transmission to many receivers and because power conservation is more essential. Consequently, the previously discussed multistation constraints do not apply to this link. Today, at least, power economy and bandwidth availability dictate trading bandwidth for power reduction, as is possible either with FM or PCM.

The former clearly is preferable because of the ease of changing SSB to FM in the satellite. In a reasonably light and reliable satellite, the problems of sampling a 5 Mc baseband 10^7 times per second and of quantizing each sample via PCM appear quite discouraging. Hence, it is a seemingly clear conclusion that FM is the most logical choice for the down link.

Initially, with FM, the deviation may be relatively high and the power correspondingly low. Later, repeater capacity can be increased (within the same RF bandwidth) by reducing the deviation and increasing the power. Except as limited by frequency-sharing problems, best spectrum utilization could be obtained with SSB. The eventual use of SSB for the down link seems problematical if only because CCIR presently recommends* that a satellite's power flux density at the earth's surface not exceed -130 dBW/m^2, nor a spectral density of -149 dBW/m^2/4 kc. Another study** under this contract is directed toward techniques for better control of interference between stationary satellites and surface microwave systems.

D. Peak Loading of FDM Voice Circuits — Ground Station Economics

A final significant conclusion resulting from this study is that the peak power required for SSB and FM is practically constant as n varies from 1 to about 150 channels. The rms power increases with n according to the channel loading, $4 \log_{10} n$, but the peak/rms loading decreases almost as fast. Bandwidth increases directly with n, but a thorough investigation of transmitter costs must be carried out to determine the total variation of transmitter costs with n. It appears to be that these costs are primarily determined by peak power rather than rms so that final ground transmitter costs are relatively fixed at the present time for $n < 150$ channels in SSB and FM systems. Reduction of transmitter costs appears to depend upon useful compandor improvement and success of depressed collector traveling-wave tubes (or similar techniques).

*Doc. 2291, Xth Plenary Assembly, CCIR, Geneva 1963.

**Report No. 7 — forthcoming.

APPENDIX A

FM IMPROVEMENT FOR FDM SYSTEMS

The problem here is that the (S/N) for each channel is given along with the number of channels, loading factor (channel and peak/rms), and the bandwidth of each channel. From these data we wish to know the (S/N) of the composite wave to be transmitted. In other words, we wish to find out how much the (S/N) per channel is reduced for the composite FDM system including the reduction for FM improvement. From refs. 16, 19, or 22 the FM improvement factor, I_{FM} is given by (1A) (with respect to SSB) for the case of f_m, the highest modulating frequency, equaling the baseband and Δf being the peak deviation for transmission of the composite baseband.

$$I_{FM} = \frac{3}{2}\left(\frac{\Delta f}{f_m}\right)^2 = \frac{3}{2}\beta^2 = \frac{\left(\frac{S}{N}\right)_{\text{FM output}}}{\left(\frac{S}{N}\right)_{\text{SSB output}}} \qquad (1A)$$

where $(\Delta f/f_m)$ is the modulation index and the same transmitter power is used for both FM and SSB. Emphasis is assumed here which provides the same $(S/N)_{ch}$ for all channels as well as the total composite baseband. Had emphasis not been used, the $(S/N)_{ch}$ for the top baseband channel would be worse by a factor of 3 than (1A) indicates. The S/N ratio for the composite FDM wave is given by (2A) where S is increased by L_{fp} and N by n.

$$\left(\frac{S}{N}\right)_c = \left(\frac{S}{N}\right)_{ch} \times \frac{L_{fp}}{n} \qquad (2A)$$

where L_{fp} = power loading factor including both channel loading and peak/rms loading. This results from the signal power being increased by the loading and the noise being assumed "white" which means it is simply proportional to bandwidth. $(S/N)_c$ is the composite (S/N) ratio in the output required for the system, FM in this case.

Transmitted S/N, C/N, is given simply by dividing (2A) by the overall improvement factor, (1A), and increasing the noise by the increased bandwidth of the transmitted FM wave.

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$$\left(\frac{C}{N}\right) = \left(\frac{S}{N}\right)_{ch} \times \frac{L_{fp}}{n} \times \frac{1}{\frac{3}{2}\left(\frac{\Delta f}{nB_{ch}}\right)^2} \times \frac{nB_{ch}}{B_{FM}} \qquad (3A)$$

where

Δf = peak deviation of the composite FM wave

nB_{ch} = highest modulating frequency

B_{FM} = over-all FM bandwidth

C/N = transmitted S/N ratio.

Since the $(\Delta f)^2$ is simply the per channel frequency deviation squared times the power loading factor,

$$\frac{C}{N} = \left(\frac{S}{N}\right)_{ch} \frac{1}{\frac{3}{2}\left(\frac{\Delta f_{ch}}{nB_{ch}}\right)^2 \frac{B_{FM}}{B_{ch}}} \qquad (4A)$$

where Δf_{ch} = per channel peak frequency deviation. Or in dB

$$\frac{C}{N} = \left(\frac{S}{N}\right)_{ch} \text{dB} - 20 \log_{10} \frac{\sqrt{3}\ \Delta f_{ch}}{nB_{ch}} - 10 \log_{10} \frac{B_{FM}}{2B_{ch}} \quad . \qquad (5A)$$

Actually, the highest modulating frequency might be higher than nB_{ch} because of CCIR channel allocation. Under this condition nB_{ch} must be changed accordingly.

Another equivalent result is obtained by using the composite S/N ratio and the over-all frequency deviation as in (6A) which follows from (3A).

$$\frac{C}{N} = \left(\frac{S}{N}\right)_c \times \frac{1}{\frac{3}{2}\left(\frac{\Delta f}{nB_{ch}}\right)^2\left(\frac{B_{FM}}{nB_{ch}}\right)} \tag{6A}$$

$$\frac{C}{N}\,dB = \left(\frac{S}{N}\right)_c dB - 20 \log_{10}\left(\sqrt{\frac{3}{2}}\ \frac{\Delta f}{nB_{ch}}\right) - \log_{10}\frac{B_{FM}}{nB_{ch}}$$

where $(S/N)_c\,dB = (S/N)_{ch}dB + (L_{fp})dB - 10 \log_{10} n$.

Actually, (6A) is probably the best formula to use in general because there are times when an over-all modulation index is assumed in the wide-band cases and no need arises to calculate the individual channel deviation. Also B_{FM} is a function of Δf and not Δf_{ch}.

Consideration is now given to FMFB (frequency modulation with feedback detection). In this case the bandwidth of the intermediate frequency filters within the detector feedback loop can be considerably reduced because of the frequency tracking action. This effectively reduces the noise power at the discriminator input while the signal or carrier power remains high. What has happened is that the wide deviation FM has been converted by the frequency tracking action to narrowband FM. As a result, the over-all FM improvement from detector input to output has remained the same as without feedback but the "effective" threshold at the detector input has been decreased. This reduction is "effective" threshold is approximately equal to the noise reduction. Exactly how much the bandwidth can be reduced is a question of practical design limitations. However, from the literature, ref. 13, a figure of three times the baseband appears possible for the i.f. bandwidth, so that value will be used here.

$$\text{Noise reduction} = \frac{B_{fm}}{3_n B_{ch}} = \frac{2}{3}\left(\frac{\Delta f}{nB_{ch}} + 1\right) . \tag{7A}$$

The threshold is thus given by

$$\frac{C}{N_t} = 12 - 10 \log_{10} \frac{2}{3}\, \beta + 1 \tag{8A}$$

where

$$\frac{C}{N_t} = \text{threshold for feedback detection}$$

$$\beta = \frac{\Delta f}{nBn}, \text{ assuming } nB_{ch} \text{ is the highest modulating frequency, } f_m .$$

If f_m is larger than nB_{ch}, then the β term must be reduced accordingly.

APPENDIX B

OVERLOAD FACTOR

Various published results show that as the number of FDM speech channels are increased the composite becomes more and more like gaussian noise. In fact, the approximation that the composite has a gaussian probability distribution is quite accurate for n, the number of channels, > 64 according to ref. 11. In that reference, experimental values are given for peak to rms ratios where the peak is exceeded 1 % and 0. 1 % of the time. A simple calculation assuming a gaussian distribution of voltages is as follows:

$$P(x) = \frac{2}{\sqrt{2\pi}\,\sigma} \int_x^{\infty} e^{-y^2/2\sigma^2}\, dy \quad . \tag{1B}$$

P (x) is the probability of $y > x$ where σ is the rms value. If we let $\sigma = 1$, the resulting value of x will be the peak voltage to rms voltage for a specified P (x).

Case (1): $$P(x) = 1\% = 0.01 = 1 - \frac{2}{\sqrt{2\pi}} \int_0^{x} e^{-y^2/2}\, dy \tag{2B}$$

and

$$\frac{1}{\sqrt{2\pi}} \int_0^{x} e^{-y^2/2}\, dy = \frac{0.99}{2} = 0.495 \quad . \tag{3B}$$

From (3B) and a table of values for the normal curve, we can solve for x.

$$x = 2.575 \tag{4B}$$

$$\frac{\text{peak power}}{\text{rms power}} = x^2 = 6.62 \text{ or } 8.21 \text{ dB} \quad . \tag{5B}$$

This agrees with Fig. 2

Case (2): $P(x) = 0.1\% = 0.001$

$$\frac{1}{\sqrt{2\pi}} \int_{o}^{x} e^{-y^2/2} \, dy = \frac{0.999}{2} = 0.4995 \qquad (6B)$$

and

$$x = 3.29$$

for which

$$\frac{\text{peak power}}{\text{rms power}} = 10.8 \text{ or } 10.34 \text{ dB} . \qquad (7B)$$

This also agrees with Fig. 2 for the average value of peak/rms ratio. In this report a value of 10.2 dB is used. A number of other papers use a value of 13 dB which is an experimental upper bound from ref. 11.

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